

HDL-TM-81-9

LEVEL

H

(D)

April 1981

AD 174 074 41  
Application of Filters for High-Altitude Electromagnetic  
Pulse Protection

by Gary L. Roffman

AD



U.S. Army Electronics Research  
and Development Command  
Harry Diamond Laboratories  
Adelphi, MD 20783

Approved for public release; distribution unlimited.

DTIC ELECTED  
S JUN 30 1981

DTIC FILE COPY

81 6 29 292

The findings in this report are not to be construed as an official Department of the Army position unless so designated by other authorized documents.

Citation of manufacturers' or trade names does not constitute an official indorsement or approval of the use thereof.

Destroy this report when it is no longer needed. Do not return it to the originator.

## UNCLASSIFIED

SECURITY CLASSIFICATION OF THIS PAGE (When Data Entered)

REPORT DOCUMENTATION PAGE		READ INSTRUCTIONS BEFORE COMPLETING FORM
1 REPORT NUMBER HDL-TM-81-9	2 GOVT ACCESSION NO AD-A100 741	3 RECIPIENT'S CATALOG NUMBER
4. TITLE (and Subtitle) Application of Filters for High-Altitude Electromagnetic Pulse Protection		5. TYPE OF REPORT & PERIOD COVERED Technical Memorandum
7. AUTHOR(s) Gary L. Roffman		6. PERFORMING ORG REPORT NUMBER
9. PERFORMING ORGANIZATION NAME AND ADDRESS Harry Diamond Laboratories 2800 Powder Mill Road Adelphi, MD 20783		10. PROGRAM ELEMENT, PROJECT, TASK AREA & WORK UNIT NUMBERS Prog Ele 33126K PRON WS9144 00WSA9 MIPR-HC1001-9-400
11. CONTROLLING OFFICE NAME AND ADDRESS Defense Communications Engineering Center 1860 Wiehle Ave Reston, VA 22090		12. REPORT DATE Apr 1981
14. MONITORING AGENCY NAME & ADDRESS (if different from Controlling Office)		13. NUMBER OF PAGES 53
15. SECURITY CLASS (of this report) UNCLASSIFIED		
15a. DECLASSIFICATION/DOWNGRADING SCHEDULE		
16. DISTRIBUTION STATEMENT (of this Report) Approved for public release, distribution unlimited		
17. DISTRIBUTION STATEMENT (of the abstract entered in Block 20, if different from Report)		
18. SUPPLEMENTARY NOTES HDL Proj E219E3		
19. KEY WORDS (Continue on reverse side if necessary and identify by block number) HEMP Filters EMP Terminal protection devices		
20. ABSTRACT (Continue on reverse side if necessary and identify by block number) The application of filters for the protection of electrical equipment from electrical transient signals induced by high-altitude electromagnetic pulse (HEMP) is discussed, and the application of filters to ac and dc power supplies and analog and digital signal inputs is described. The application of filters is demonstrated for different types of filters, also shown are filter limitations caused by arcing, source and load impedance mismatches, saturation, and parasitic reactances. The use of filters following a nonlinear terminal protection device (TPD) is shown to be a very effective means of transient energy reduction. Magnetic cores and ferrite beads can provide a simple method of high-frequency transient suppression		

CONTENTS

Page

1. INTRODUCTION .....	7
2. TYPES OF FILTERS .....	8
3. FILTER PRACTICE APPLICATIONS .....	11
3.1 Alternating Current Power Lines.....	11
3.2 Direct Current Power System.....	21
3.3 Analog Signals .....	30
3.3.1 Cable Shield Coupling Reduction with Magnetic Cores.....	31
3.3.2 Ferrite Shield Beads .....	33
3.3.3 Miniature Pin Filters .....	37
3.3.4 Ferrite Beads and Nonlinear TPDs .....	38
3.3.5 Balanced Analog Signal Transmission .....	40
3.3.6 Crystal, Ceramic, and Mechanical Filters .....	42
3.3.7 Problems with Analog Filters.....	43
3.4 Digital Circuit Filtering.....	45
3.4.1 Balanced Digital Signal Transmission.....	46
3.4.2 Balanced Circuit Example .....	49

FIGURES

1 Low-pass filter sections: (a) Response of ideal low-pass filter, (b) ladder-type filter, and (c) $\pi$ and Tee filter sections .....	9
2 Component value determination for low-pass filter sections: (a) Component values for constant-K Tee filter, (b) component values for constant-K $\pi$ filter, (c) component values for m-derived, Tee filter, (d) component values for m-derived, $\pi$ filter .....	9
3 Ratio of power out to power in of a lossless and dissipative filter.....	11
4 Power service entrance treatment (a) Rectified power supply, (b) filter at zone $1/2$ interface barrier for two-wire power line, and (c) grounding of green wire for three-wire power service.....	12
5 Dissipative power-line filter schematic and insertion loss.....	14
6 Lossless power-line filter schematic and insertion loss.....	15

FIGURES (Cont'd)

	<u>Page</u>
7 Schematic for power ratio in decibels with mismatched source and load impedance	15
8 Frequency response of filter in figure 6 with different source and load impedances	16
9 Schematic of filter and frequency response of filter with different source and load impedances	17
10 Dissipative filters: (a) resonance reduction using dissipative filters and (b) construction of commercial filter	18
11 Schematic of dc power supply for communication facility	22
12 Uninterruptible power supply for commercial facility	22
13 Power supply isolation using dc-dc converter	23
14 Specification and insertion loss of commercial filters	25
15 Simple RC filter	25
16 Simple LR filter	26
17 Simple LRC filter	26
18 Impedance of ferrite bead with different dc currents	27
19 Insertion loss of EMI suppressant tubing and ferrite beads	28
20 Simple filters for common-mode suppression (a) bifilar (top) and standard (bottom) wound transformers and (b) nonsaturating dc power-line filter	29
21 Reducing pickup on small-signal cable lines	30
22 Communication cable using choke cores and multiple grounds to eliminate shield currents	32
23 Ferrite bead on wire and ferrite bead equivalent circuit	33
24 Commonly available sizes of ferrite beads	33
25 Ferrite bead frequency-dependent resistance and reactance for two ferrite materials	34
26 Circuit for estimating insertion loss	34

FIGURES (Cont'd)

	<u>Page</u>
27 Equivalent circuit used with NET-2 to simulate ferrite bead in an electronic circuit	35
28 Measured inductance and resistance of ferrite bead	36
29 Response of ferrite bead simulated with NET-2 circuit-analysis program compared to low-frequency inductance model of bead	36
30 Attenuation with large numbers of ferrite beads	37
31 Attenuation of miniature pin filters	37
32 EMP-induced sinusoidal signal at top passes through three different TPDs	39
33 Peak output voltage for a TPD and TPD filter combination	39
34 Effect of capacitor ( $C_F$ ) lead length on filter performance	40
35 Balanced line isolation using transformer	41
36 Transformer and equivalent circuit of transformer	42
37 (a) Primary tuned transformer and (b) secondary tuned transformer	42
38 Fourier transform of a pulse showing improved approximation to pulse as frequency is increased.	46
39 Normalized bandwidth of an EMP signal with an exponential rise and decay	46
40 Shunt and series transformer wiring configuration	47
41 Common-mode driving of series transformer	47
42 Common-mode rejection test of shunt and series transformer configuration with standard and bifilar windings	48
43 Differential output of transformers with shunt and series configuration and bifilar and standard windings	49
44 No. 1 ESS pulse transmission system	49

## TABLES

	<u>Page</u>
1 Filters That Survived an 11-kV, 50-ns Pulse Test Performed at the Harry Diamond Laboratories	19
2 Filter Manufacturing Companies	20
3 Filter Cost Versus Power Handling Capability	21
4 Occurrence of Arcing at Various Voltages	38
5 Energy Reduction Produced by TPD Ferrite Bead Filter Combination	40
6 Summary of Filter Frequency Range, Advantages, and Disadvantages	44
 LITERATURE CITED	50
 DISTRIBUTION	53

## 1 INTRODUCTION

This paper describes the application of filters to protect the inputs to electrical equipment against electrical transients produced by the electromagnetic pulse (EMP) associated with a nuclear detonation. Various types of electromagnetic interference (EMI) filters may already be used in connection with equipment signal and power supply inputs. The complete design of such filters is beyond the scope of this report, which is to describe the application of filters for EMP protection. The problem with many commercial filters is that the high rise-time, high-amplitude signals produced by an EMP event may cause voltage breakdown or high-frequency transmission, which can nullify the filter function. The design of filters for high-voltage and high-frequency use is critical to EMP filter requirements.

Filters are useful for protection against EMP or other types of EMI when the filter passband corresponds to the frequency range (on dc) of desired signals and the stop band corresponds to the undesired signal. The energy in an EMP signal exists over a wide frequency range, but in general a low-pass or bandpass filter can reduce the energy induced by an EMP signal to levels below the damage threshold and often below the upset threshold of the equipment protected. A problem in some types of circuits—for example, digital circuits—is that the signal frequency range and the disturbance frequency range can overlap. Since disturbance signals are usually coupled to wires or cables as a large common-mode signal, isolation transformers and bifilar chokes can be used to suppress the common-mode disturbance signal but pass the differential signal. Microwave frequencies are higher than the spectrum produced by an EMP signal, so distributed bandpass and high-pass filters can be used to protect microwave systems.

The decision as to what type of terminal protection device (TPD) to use for a particular application depends on the following considerations:

- a. Bandwidth of the signal relative to the EMP signal spectrum
- b. Magnitude of the disturbance suppression required
- c. Reliability or damage vulnerability of the TPD fix
- d. Inadequate nonlinear TPD fixes, i.e., low-amplitude signals or transient disturbance pulses bypassing other types of TPD fixes<sup>1</sup>
- e. Need for large currents, as in power supplies
- f. Cost of the TPD

The effectiveness of a filter compared to nonlinear TPD's (spark gap, varistor, or Transzorb) is greatest when the bandwidth of the signal is outside the EMP spectrum (0.1 to 100 MHz) or occupies only a small portion of the EMP spectrum, since the filter discriminates by frequency rather than amplitude. A filter will often have a limited capability for attenuation, either because of the design of the filter or because of parasitic reactances associated with the filter components. So the amount of disturbance attenuation required to prevent damage has to be considered. Since the filters are composed of passive components, the filter components are generally more

<sup>1</sup>E. F. Vance et al., *Technical Inputs and EMP Design Practices for Intrasite Cabling of Telecommunication Facilities*, Stanford Research Institute, prepared for HDL under contract DAAG39-76-C-0021 (June 1977), 109.

damage resistant than a nonlinear semiconductor TPD. Therefore, a filter is less likely to be damaged by an EMP-induced signal than such a nonlinear TPD and can continue to protect a circuit in a multiple burst environment.<sup>2</sup> In fact, filters can be composed of self-healing elements, if indeed transient levels exceed rated thresholds.

Linear components are generally less vulnerable than the nonlinear semiconductor components to changes in characteristics from aging, nuclear radiation, or thermal effects. Because filters are not amplitude dependent, they can be used with low-signal-level digital devices where adequate protection cannot be provided by a nonlinear TPD because the protection threshold is too high. The use of both a nonlinear and linear TPD for the protection of low-signal-level digital devices may be required where a filter alone cannot adequately suppress the disturbance signal. Filters can also complement the function of a nonlinear TPD such as a spark gap by filtering the transient pulses which can bypass the nonlinear TPD due to its finite turn-on time. Filters can be designed to pass large currents without affecting their frequency response, which is useful in protecting dc and low-frequency ac supplies. However, inductors with ferrite cores are subject to saturation effects. Because filters are often used in circuits for EMI protection, there may often be no cost in implementing filter protection other than determining whether the filters are adequate for EMP-induced signal protection. The EMP cumulative energy-density spectrum shows that most energy is above 100 kHz, so that relatively low-value inductors and capacitors can be used for filtering. Low-value components are generally small and inexpensive. However, components that can withstand high voltages are required, which can be more expensive. The cost of filters will be discussed more thoroughly in the subsections concerning particular filter types.

## 2 TYPES OF FILTERS

Figure 1(a) shows the response of an ideal low-pass filter. At frequencies below the cutoff frequency no power is lost, and at frequencies above the cutoff frequency all power is lost. An actual filter differs from the ideal by having attenuation in the passband and a finite slope in the attenuation above the cutoff frequency. Different types of filters are synthesized to approximate the characteristics of the ideal filter. Ladder-type filters (fig. 1(b)) are often used to approximate the ideal filter characteristic. The impedance into the filter is  $Z_{11}$  with  $Z_S$  disconnected and  $Z_L$  connected. The impedance out of the filter is  $Z_{22}$  with  $Z_S$  connected and  $Z_L$  disconnected.<sup>1</sup>

$$Z_S = Z_{11} \text{ and } Z_L = Z_{22} \quad (1)$$

The filter is matched to the source and load. The design of many filters is based on the impedance match in equation 1, and the filter response is affected by impedance mismatches. The basic ladder filter elements are  $\pi$  and Tee sections (see fig. 1(c)). The Tee section is often preferred for EMP hardening because with the  $\pi$  section high voltages can develop across the input capacitor and cause the capacitor to degrade or fail due to a resonance between the capacitor and source.<sup>3</sup> The Tee section can withstand these high voltages, but more power may be delivered to the load.

<sup>1</sup>IIIT Research Institute DNA EMP Awareness Course Notes, DASA 01-69-C 0095, Defense Nuclear Agency, DNA 2772T (August 1974), 41

<sup>2</sup>R. A. Perala and T. F. Ezell, *Engineering Design Guidelines for EMP Hardening of Naval Missiles and Airplanes*, Mission Research Corp., prepared for the Naval Ordnance Laboratories under contract N60921-73-C-0033 (December 1973) Ch 7, 3

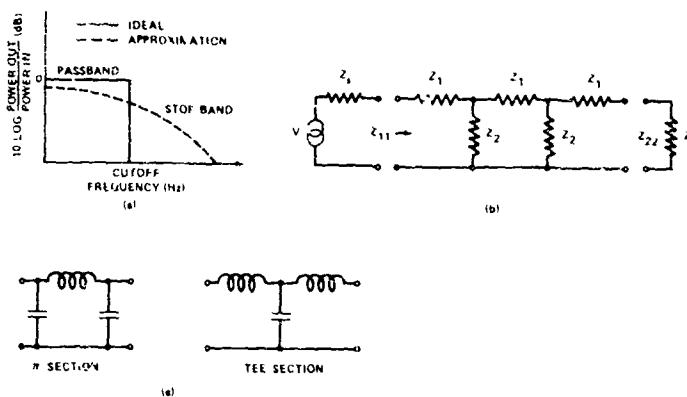


Figure 1 Low-pass filter sections (a) Response of ideal low-pass filter, (b) ladder-type filter, and (c)  $\pi$  and Tee filter sections

If  $Z_1 Z_2 = Z_L^2$  in figure 1(b), the constant-K class of filters is defined. Figures 2(a) and 2(b) show the values of the components for a Tee and  $\pi$  section for a constant-K filter. For a constant-K Tee section, if a new series impedance  $Z_1'$  is defined, where  $Z_1' = m Z_1$  ( $Z_1$  is the old series impedance), a new value for  $Z_2$  is required if the impedance looking into and out of the filter ( $Z_{11}$  and  $Z_{22}$ ) is to remain the same as with the constant-K filter. Figures 2(c) and 2(d) show the components that produce the new value of  $Z_2$  for Tee and  $\pi$  sections that are required if the impedance looking into and out of the filter ( $Z_{11}$  and  $Z_{22}$ ) is to remain the same as with the constant-K filter.<sup>4</sup> This type of filter is called an m-derived filter. Note that if  $m = 1$ , figures 2(a) and 2(b) are identical to figures 2(c) and 2(d), respectively. The value of these types of filters is that, for a given load and source resistance, the basic  $\pi$  or Tee sections can be staged, and the impedance between the stages can be matched. The staged sections will have a more ideal filter characteristic. Also, the two types of filters can be combined because of the matched impedance characteristic. Similar circuits and formulas for component values can be derived for high-pass and bandpass filters.<sup>5</sup>

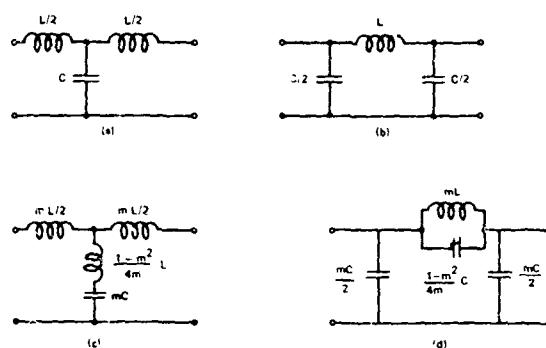


Figure 2 Component value determination for low-pass filter sections  
 (a) Component values for constant-K, Tee filter; (b) component values for constant-K,  $\pi$  filter  
 (c) component values for m-derived, Tee filter, (d) component values for m-derived,  $\pi$  filter

<sup>4</sup>M. E. Van Valkenburg, *Network Analysis*, Prentice Hall, Inc. (1955), Ch 13

<sup>5</sup>Donald G. Fink, *Standard Handbook for Electrical Engineers*, 10th edition, McGraw-Hill (1969), sect. 2

The attenuation characteristics of the constant-K filter near the cutoff frequency are not as sharp as for the m-derived filter, so more stages of constant-K filter sections and thus more components would be required for the same attenuation rate. Note however that each section (fig 2(b)) of the m-derived filter contains an extra component. Another possible advantage of the m-derived filter is that the impedances into and out of the filter remain relatively constant in the passband, especially if  $m = 0.6$ .<sup>4</sup> Thus, if the load and source impedances are constant (resistive), a better impedance match occurs.

If a flat frequency response is desired in the passband, Butterworth and Tchebycheff filters can be used. The Tchebycheff filter gives an optimum rate of attenuation after the cutoff frequency, but is much more critical to component value variation.<sup>5</sup> The basic  $\pi$  and Tee sections for the Butterworth and Tchebycheff filters are the same as the constant-K filter (fig 1(c)). Tables of component values for these filters exist with equal load and source resistances.<sup>6</sup> If the Q of the inductors and capacitors comprising the filters is considered, then different component values are required. Tables are given of component values for filters with lossy inductors or lossy inductors and capacitors.<sup>6</sup> The Q of the reactive components can be adjusted by placing resistors in series with inductors and in parallel with capacitors. Filters with low Q are considered to be dissipative filters and will be discussed later. Another use for dissipative filters is that they are singly loaded, i.e., either a resistive load or a resistive source impedance is required, but not both. Therefore, the source impedance can be zero with a resistive load or the load impedance can be infinite with a resistive source. These load conditions can be approximated by a load impedance relatively large compared to the source impedance and the converse.<sup>6</sup> Lossless filters can also be designed to function between different source and load impedances.<sup>6</sup> The Butterworth and Tchebycheff low-pass filters can be transformed to produce high-pass<sup>6</sup> and bandpass<sup>6</sup> filters.

A dissipative filter absorbs energy rather than reflecting energy. This type of filter is generally preferred for use with EMP-induced signals, since the location at which energy is dissipated is controlled. With a reflective filter, energy is dissipated somewhere else in the system where it may cause damage. Dissipative filters can occur because of the use of low-Q inductors or capacitors or can be deliberately designed by adding resistors in series with inductors or in parallel with capacitors. The effects of dissipation on a lossless filter are that a fixed loss over the entire spectrum is introduced, load and source impedances are not as critical,<sup>3</sup> ripples in the passband are smoothed out (Tchebycheff), the attenuation near the cutoff frequency is rounded off, and infinite attenuation becomes finite.<sup>6</sup> Figure 3 shows an example of the effects of dissipation on a lossless filter design with equal source and load impedance. The use of ferrite beads or torroids which have an RL equivalent circuit provides a convenient method of producing dissipation. The use of ferrite materials and dissipative filters will be discussed in more detail later.

---

<sup>3</sup>R. A. Perala and T. F. Ezell, *Engineering Design Guidelines for EMP Hardening of Naval Missiles and Airplanes*, Mission Research Corp., prepared for the Naval Ordnance Laboratories under contract N60921 73-C-0033 (December 1973), Ch 7, 4.

<sup>4</sup>M. E. Van Valkenburg, *Network Analysis*, Prentice Hall, Inc. (1954), 346.

<sup>5</sup>Philip R. Geffe, *Simplified Modern Filter Design*, Hayden Book Company, Inc. (1963), 2, app 2, app 3, 17, 18, Ch 2, Ch 3 and 4, and p 11.

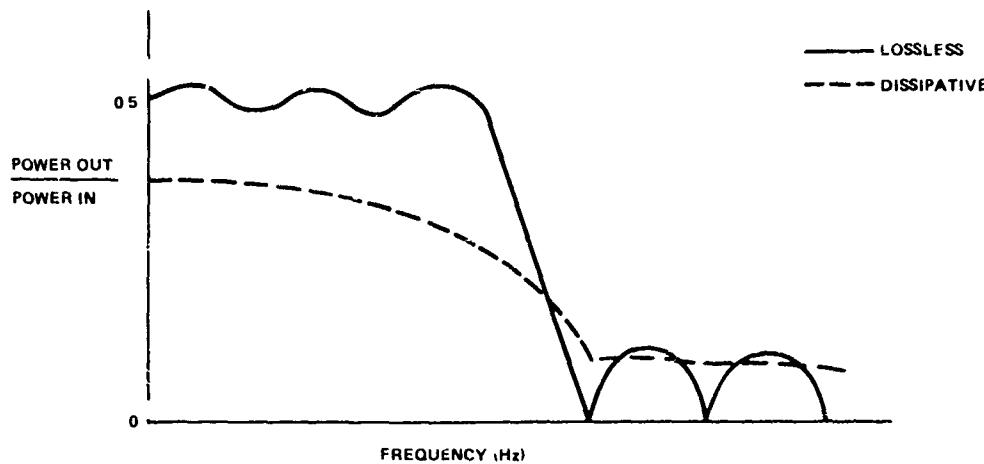


Figure 3. Ratio of power out to power in of a lossless and dissipative filter

### 3. FILTER PRACTICE APPLICATIONS

The following sections will discuss in detail methods of hardening the various types of lines or cables connected to equipment. Included are ac and dc power lines, grounds, and signal lines with either analog or digital circuits. The filter designs previously described require fixed load and source impedances and the repeated use of certain simple circuit configurations ( $\pi$  or Tee). This allows for the design of filters with some desired frequency characteristic. Commonly used types of filters have been briefly described mainly to provide familiarization with the type of filters that are commercially available.

An important consideration when adding a filter to a circuit is the effect of the filter on normal circuit operation. Also, the source and load requirements of the filter will affect its frequency characteristics. Computer circuit-analysis codes<sup>7-10</sup> are valuable tools in predicting the effects of the insertion of a filter on normal circuit operation. These codes are also helpful in predicting how the use of mismatched source and load impedances will affect filter performance. In addition, with these computer codes it is possible to model the effects of nonlinear source and load impedances. Examples will be shown of the use of computer circuit-analysis codes for hardening circuits with filters.

#### 3.1 Alternating Current Power Lines

For most communication facilities, the power required for day-to-day operation is derived from commercial utility sources. Power from these sources is used for essential com-

<sup>7</sup>Allan F. Malmberg, NET-2 Network Analysis Program Release 9, BDM Corporation, prepared for HDL under contract DAAG39-70-C-0050, HDL 050-1 (September 1973).

<sup>8</sup>Laurence Nagel, SPICE-2: A Computer Program to Simulate Semiconductor Circuits, Electron Research Laboratory, College of Engineering, University of California at Berkeley (May 1976).

<sup>9</sup>L. D. Milliman et al, CIRCUIS: A Digital Analysis Program of Electronic Circuits—Program Manual, the Boeing Corporation, prepared for HDL under contract DA-49-186-AMC-346(X), Harry Diamond Laboratories, HDL-346-2 (January 1967).

<sup>10</sup>BDM Corporation, NET-2 Network Program Analysis Addendum to User's Manual for Release 9.1, prepared for HDL under contract DAAG39-77-C-0150, BDM/W-77-573-TR (November 1977), Ch 5.

unication equipment as well as for expendable functions such as lighting, personnel accommodations, and other purposes not essential to short-term communication. Because commercial power is subject to occasional failure—particularly during such emergencies as national disasters—most permanent communication facilities also contain emergency generators of sufficient capacity to carry the essential equipment loads. The facilities may store enough fuel to supply power for periods ranging from a few days to more than a month.<sup>1</sup>

From the standpoint of EMP hardening, therefore, the primary concern is not preserving the external source of ac power, but rather controlling the EMP-induced interference carried into the facility on power conductors.\* The overhead utility distribution lines outside the building form a very large collector of EMP energy and serve to guide this energy to the facility along the lines serving the facility. The distribution transformer provides some isolation of the distribution lines from the low-voltage wiring in the facility, however, coupling through the transformer is fairly efficient in the frequency range between 100 kHz and 10 MHz. Unfortunately, many of the natural resonances of the communication facility are in the passband of the distribution transformer.

By filtering the power lines between the outside and the inside of the facility with filters providing at least 60 dB (80 dB preferred) of attenuation at frequencies above 100 kHz<sup>1</sup> and by good shielding of the building, the EMP-induced signal level inside a facility can be greatly reduced. However, if proper and consistent filtering and shielding is not provided, signal levels capable of damage or upset can get to the equipment cabinets and into the equipment. Equipment manufacturers often provide some type of ac power-line filtering as part of the rectifier/power supply (fig. 4(a)). This filtering may not be adequate, and additional filters may be required. Figure 4(b) shows the placement of filters for a two-wire single-phase power input.<sup>1</sup> Note that both wires are filtered because of the possibility of a large common-mode signal. If a three-wire service (black, white, and green) is used, the green wire should be connected to the cabinet immediately after it enters the cabinet (fig. 4(c)).

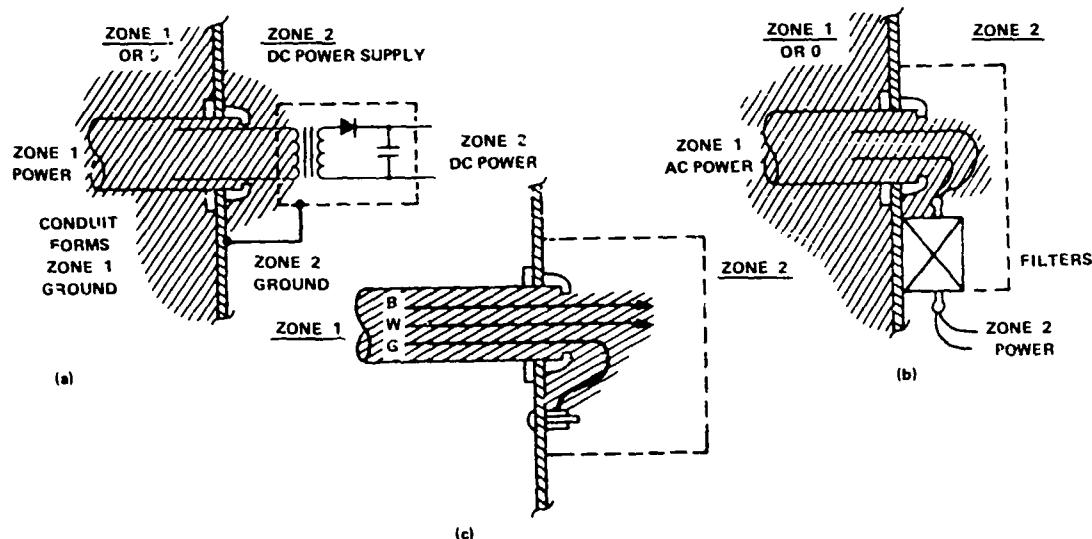


Figure 4 Power service entrance treatment (a) rectified power supply, (b) filter at zone 1/2 interface barrier for two-wire power line, and (c) grounding of green wire for three-wire power service

<sup>1</sup>E. F. Vance et al., *Technical Inputs and EMP Design Practices for Intrasite Cabling of Telecommunication Facilities*, Stanford Research Institute, prepared for HDL under contract DAAG39-76-C-0021 (June 1977)

\*Reliability of the emergency power generator and switching system in the EMP environment is an important consideration, however.

Alternating-current power filters—when required for protection of equipment connected to 60 Hz, 115 Vac power—should be located on interior shielded enclosure bulkheads (fig 4(b)) at the point of conductor penetration. Filters may not be required on very insensitive electrical equipment whose power conductors are shielded in ferrous conduit all the way from the generator to the equipment.

Alternating-current power-filter response curves measured as described in MIL-STD-220A should have a rising slope to greater than +3 dB at 1kHz with no negative values. This will help prevent voltage gain when the filter is installed in a system with impedance mismatches. Alternating current EMI power filters normally are expected to exceed 100 dB attenuation from 14 kHz to 1 GHz.<sup>11</sup> The following are some desirable characteristics of ac power filters to be used where large transient voltage pulses are expected.

a Filters should employ inductive inputs and capacitive outputs. The inductor assures firing potential for any preceding arrester and limits the current through the filter capacitor. The input inductor in these filters should be single-sweep wound and employ adequate spacing between start and finish leads to withstand at least a 5000-V transient breakdown test. The inductor core should be taped with suitable insulating material (preferably Teflon tape) to withstand at least 5000 V from the winding to the core. Cores should be of powdered iron or molybdenum permalloy. The inductor windings should employ only heavy Formvar or heavy Seldereze wire.

b The capacitors in these filters should be of self-healing Mylar or metallized Mylar and should have a breakdown voltage of not less than 600 Vdc.

c Potting material should be high-temperature (125 C) wax or neat-conductive epoxy. Oil (except for flame-retardant silicon oils) or foam material should not be used.<sup>11</sup>

Figure 5 shows the schematic of a commercial filter designed to provide 100-dB attenuation from 14 kHz to 10 GHz when used in a power circuit with 0.5 ohm or less source impedance and a load drawing 25 A.<sup>11</sup> (The first four components on the left in fig 5 represent the full power-line source impedance.) The original lossless filter in figure 5 was modified by the addition of the 10-ohm resistor and 211- $\mu$ H inductor components placed in parallel with the three original 250- $\mu$ H inductors. This modification converts the filter to a dissipative filter. The load impedance of 4.7 ohms provides a 25-A load on the filter, assuming an approximately 120 Vrms voltage source. The effects of load mismatches and energy dissipation can be seen from this example. In figure 5 the dashed curve shows the calculated response of the unmodified filter. Note the ripples and rapid rate of attenuation at frequencies above the cutoff frequency. The solid curve shows the calculated response of the modified filter. Note the great reduction of the ripples and the rounding of the attenuation curve. The measured response, however, shows a more rapid rolloff than the calculated response. The insertion loss in figure 5 shows a gain at some frequencies. This is a voltage gain produced because of the source and load impedance mismatch and is not a power gain.

<sup>11</sup>H. A. Lasitter et al, *Nuclear Electromagnetic Pulse Protection Measures Applied to a Typical Communication Shelter*, Naval Civil Engineering Laboratory (April 1970) AD 707-696, 73, 108

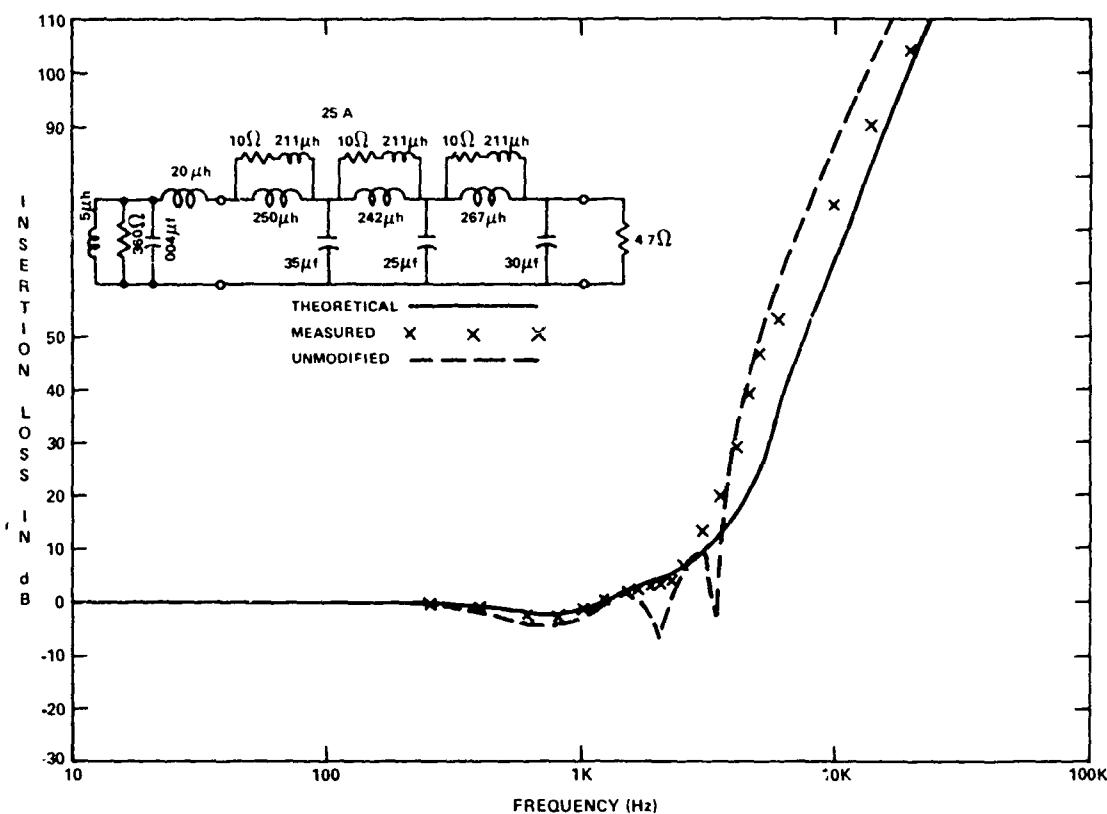


Figure 5 Dissipative power-line filter schematic and insertion loss

Figure 6 shows another filter which is lossless. The same signal source impedance is used as in figure 5. The load resistance is increased to 58.5 ohms to provide a 2-A current through the filter. Note that the frequency-response resonances (ripples) are much greater than with the dissipative filter (fig. 5). Also shown in figure 6 is the filter response with a 100-A current. The cutoff frequency is increased and resonances occur at different frequencies. The ac filter requirement of 100-dB attenuation above 14 kHz is not obtained at high current levels because the inductor cores are saturated, which reduces their reactance. However, the attenuation is greater than 80 dB at frequencies above 100 kHz.<sup>1</sup> Another effect on the frequency response is the matching of the load and source impedance to the filter. This filter was designed for a 50-ohm load and for source impedances; the mismatch causes the large amplitude resonances and the almost 30 dB of voltage gain. Since power attenuation is the critical term of EMP-induced damage, unless matched or known impedances are used, voltage attenuation can be misleading. The power ratio in decibels is (fig. 7)

$$\frac{\text{Power out}}{\text{Power in}} = 20 \log \frac{V_{out}}{V_{in}} + 10 \log \frac{R_{in}}{R_{out}} \quad (2)$$

Since  $R_{in}$  is less than  $R_{out}$  for the filter in figure 6, an additional power attenuation actually occurs

<sup>1</sup>E. F. Vance et al, Technical Inputs and EMP Design Practices for Intrasite Cabling of Telecommunication Facilities, Stanford Research Institute, prepared for HDL under contract DAAG39-76-C-0021 (June 1977), 110

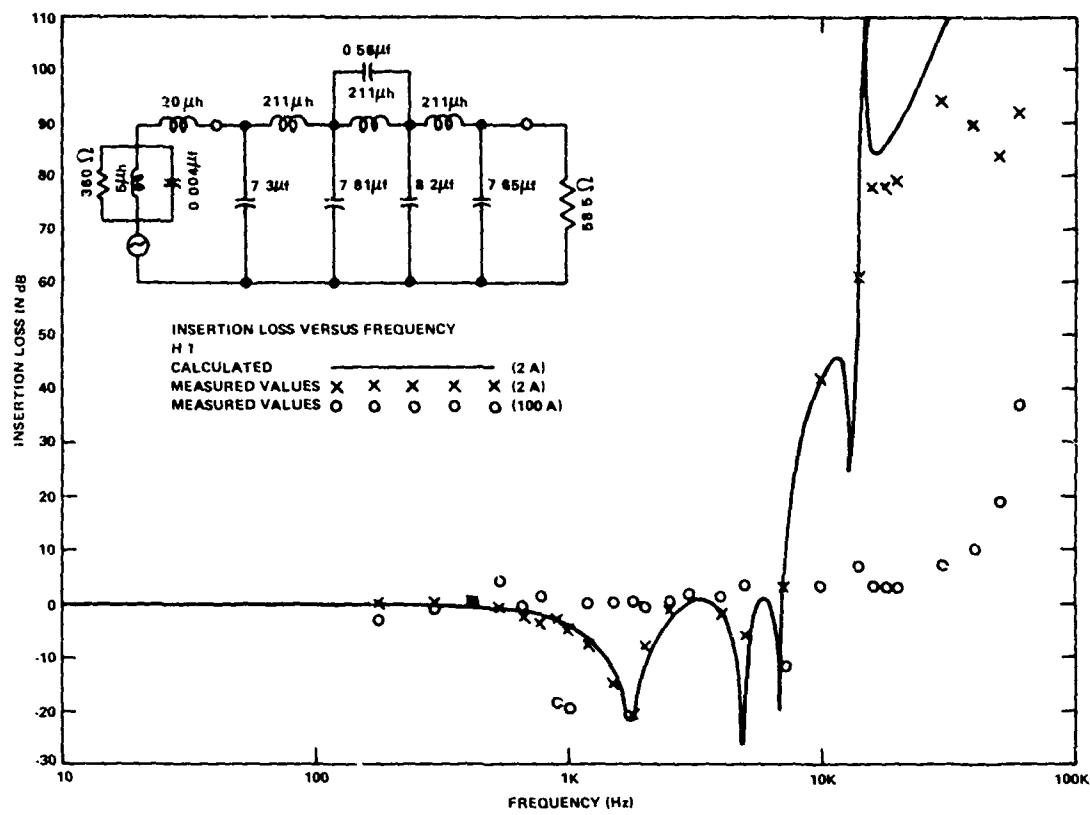


Figure 6 Lossless power-line filter schematic and insertion loss

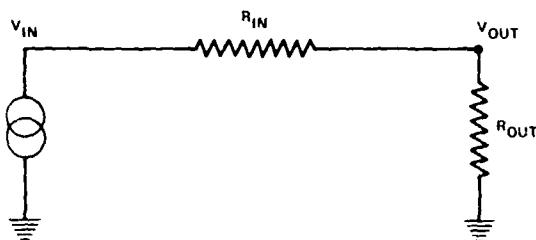


Figure 7 Schematic for power ratio in decibels with mismatched source and load impedance

Figure 8 shows the response of the same filter in figure 6, but with different source and load impedances. Using equal 50-ohm source and load impedances, the voltage gain seen in figure 6 does not occur. Also, increasing the load impedance to 200 ohms has a negligible effect on the frequency response. Reversing the load and source impedances of the last example so that the load impedance is 50 ohms and the source impedance is 200 ohms causes an overall attenuation increase, but the resonances in the frequency response still occur at the same frequencies. Increasing both the load and source impedance produces almost the same frequency response as a 50-ohm load and 200-ohm source; however, the resonances (ripples) are larger

With a multisection filter, such as the filter in figure 6, the variation in load and source impedance produces an overall constant attenuation or change in the magnitude of the ripple but does not drastically change the shape of the attenuated frequency response or the frequencies where resonances occur.

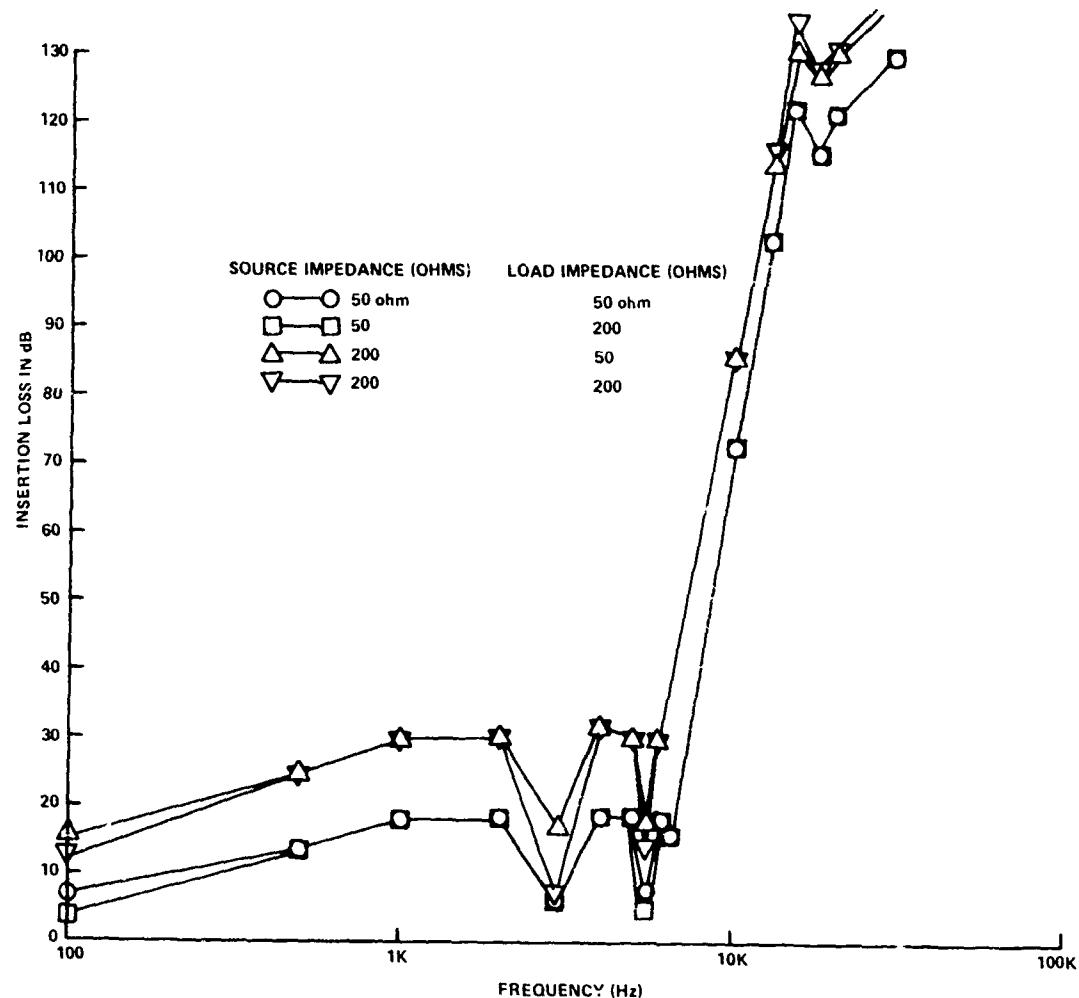


Figure 8 Frequency response of filter in figure 6 with different source and load impedances

Figure 9 shows the circuitry and frequency response of a less complex filter than the filter in figure 6. The filter is a three-pole Butterworth filter designed for a cutoff frequency of 10 kHz with 50-ohm load and source impedances. This filter is used to demonstrate loading effects and may not be useful as an ac filter because of the large inductive values. The effect of changing the load and source impedances is again predominantly to shift the frequency response by a constant amount. However, with a zero source impedance, a positive insertion loss still occurs. Increasing both the load and source impedances to 200 ohms produces more attenuation at lower frequencies and less attenuation at high frequencies. This simple filter does not produce

the rapid rate of attenuation of the multistage filter. At 14 kHz, the attenuation is 25 dB compared to over 100 dB for the multistage filter. The attenuation of the simpler filter may be adequate if the ac lines are filtered as they enter the facility and the building is reasonably well shielded. The load and source impedances used in figures 8 and 9 are purely resistive. If the load and source impedances had reactive components, resonances in the frequency response can occur which are more sensitive to the load and source impedances.<sup>3</sup> Dissipative filters would be less sensitive to load and source impedance variations.<sup>3</sup> The dependence of a particular filter with different load and source impedances can be easily calculated using circuit analysis codes<sup>7-10</sup> Unfortunately, both the attenuation and phase shift are required, and usually only attenuation is given.

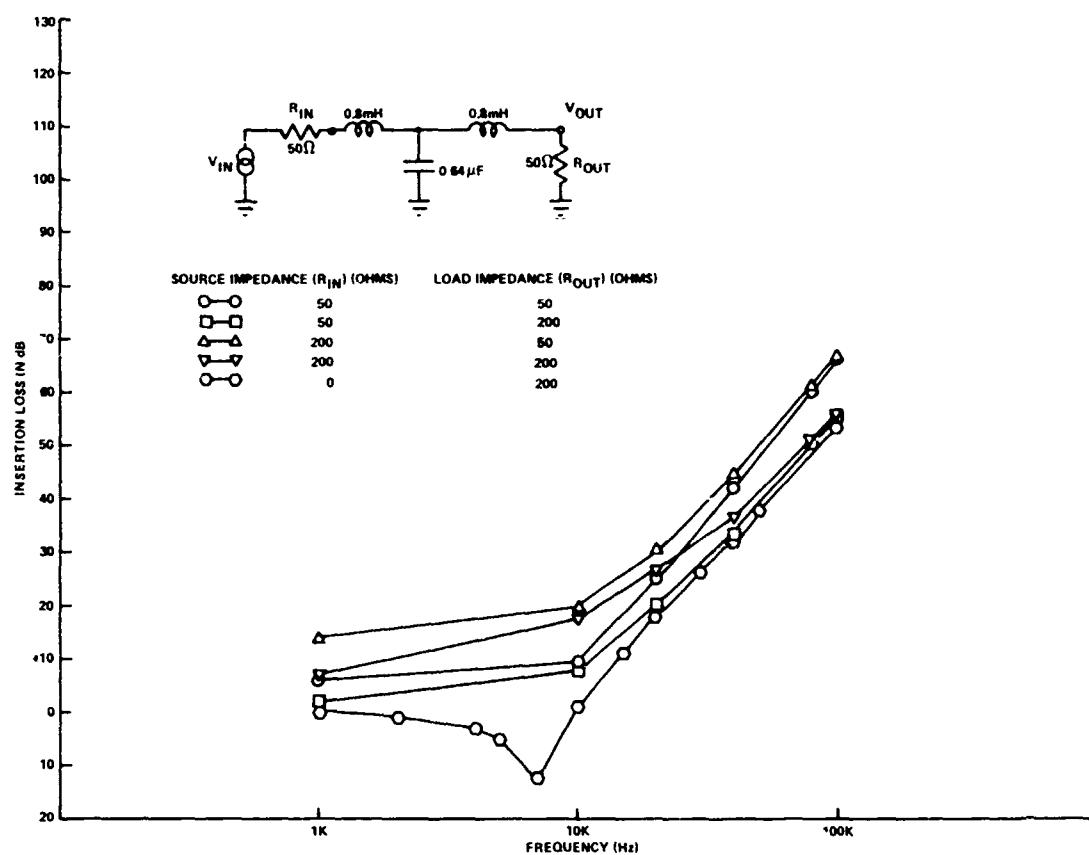


Figure 9 Schematic of filter and frequency response of filter with different source and load impedances

<sup>3</sup>R. A. Perala and T. F. Ezell, *Engineering Design Guidelines for EMP Hardening of Naval Missiles and Airplanes*, Mission Research Corp., prepared for the Naval Ordnance Laboratories under contract N60921-73-C-0033 (December 1973), Ch 7, pp 3 and 4

<sup>7</sup>Alan F. Malmberg, *NET-2 Network Analysis Program Release 9*, BDM Corporation, prepared for HDL under contract DAAG39-70-C-0050, HDL-050-1 (September 1973)

<sup>8</sup>Laurence Nagel, *SPICE-2: A Computer Program to Simulate Semiconductor Circuits*, Electron Research Laboratory, College of Engineering, University of California at Berkeley (May 1976)

<sup>9</sup>L. D. Milliman et al, *CIRCUS: A Digital Analysis Program of Electronic Circuits—Program Manual*, The Boeing Corporation, prepared for HDL under contract DA-49-186-AMC-346 (X), Harry Diamond Laboratories, HDL-346-2 (January 1967)

<sup>10</sup>BDM Corporation, *NET-2 Network Program Analysis Addendum to User's Manual for Release 9.1*, prepared for HDL under contract DAAG39-77-C-0150, BDM/W-77-573-TR (November 1977)

The voltage gain seen in figures 5, 6, and 9 occurs when the source impedance is very small compared to the load impedance. At low frequencies the source impedance of power lines is near zero.<sup>12</sup> The standard 50-ohm load and source impedances used by MIL-STD-220A to test most filters fails to show these resonances. For example, in figure 8 where a variety of source and load impedances are used, a voltage gain was not observed, but in figure 6 the same filter with a small source impedance produced voltage gain (power gain does not occur of course). The problem with these voltage gains is that the filter rings at these frequencies when subjected to a voltage pulse. Most of the energy in an EMP-induced disturbance signal is at much higher frequencies than the resonance frequencies (5 to 10 kHz) exhibited by these filters, so component damage is unlikely to occur. Other resonances due to reactive load and source impedances or filter components' parasitic reactances can cause resonances at higher frequencies. The testing of filters using MIL-STD-220A may not show these resonances because of the matched 50-ohm load and source impedances used for this test. What is required in order to determine whether a filter has resonances is to know the filter load and source impedance characteristics as well as the characteristics of the filter. If this information is not available, the use of dissipative filters can reduce resonance problems (fig 10(a)). Also, the source impedance of power lines increases at high frequencies where most of the EMP energy occurs.

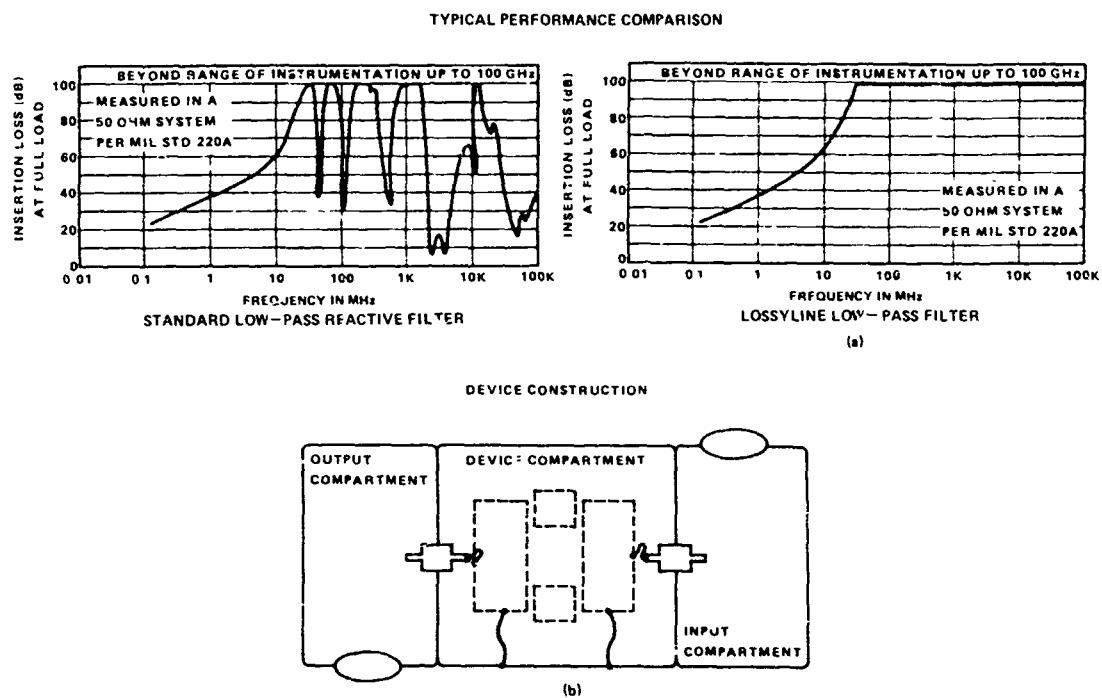


Figure 10 Dissipative filters: (a) resonance reduction using dissipative filters and (b) construction of commercial filter

<sup>12</sup>Donald L Chaffee and D B Clark, *Insertion Loss and Pulse Response of Power Line Filters*, Naval Civil Engineering Laboratory, Technical Note N-1111 (June 1970), 2, 3, 12

Commercially available dissipative filters (lossy) are available, and some have been tested with high-voltage pulses. Some EMI filters were tested by HDL<sup>13</sup> (see also table 1) and had failure levels above 11 kV (tested only to 11 kV). The filters tested are rated for high currents at dc and at 60 and 400 Hz. Most of these filters provide greater than 60-dB attenuation<sup>1</sup> at 100 kHz<sup>14</sup>. Other EMI and RFI (radio frequency interference) filters are available commercially which can provide greater attenuation and can be used with high currents at low frequencies<sup>14-15</sup> (see also table 2).<sup>16</sup> The cost of filters generally increases with power handling capacity. This relationship can be seen by the cost of the filter listed in table 3.

TABLE 1 FILTERS THAT SURVIVED AN 11 kV, 50-ns PULSE TEST PERFORMED AT THE HARRY DIAMOND LABORATORIES

Description of filter	Manufacturer	Type	$V_{bd}/I^2$
EMI	Spectrum Control	51-714-005	200 V/10 A
EMI	Spectrum Control	51-714-007	100 V/10 A
EMI	Spectrum Control	51-301-030	050 V/10 A
EMI	Spectrum Control	51-715-001	750 V/25 A
EMI	Spectrum Control	51-702-003	500 V/25 A
EMI	CAPCON	A60-B/60 Hz	125 V/60 A
EMI	CAPCON	A60-B/400 Hz	125 V/60 A
EMI	CAPCON	A10-B/60 Hz	125 V/10 A
EMI	CAPCON	A20-B/400 Hz	125 V/10 A
EMI	CAPCON	A2-B/60 Hz	125 V/2 A
EMI	CAPCON	A2-B/400 Hz	125 V/1.5 A
Lossy	CAPCON	S005-05K-155	500 V/5 A
RFI/EMI	RTRON	RNC-111	400 V/5 A
RFI/EMI	RTRON	RNC-124	400 V/2 A
RFI/EMI	U S Cap	5240-009	500 V/5 A
RFI/EMI	U S Cap	2100-026	100 V/10 A
RFI/EMI	U S Cap	2100-0263	100 V/10 A
Bandpass	Texscan	6BE-41 5/23-II	—
Bandpass	Texscan	6BD-41 5/23-II	—
Bandpass	Texscan	6BE-64 5/23-II	—
Bandpass	Texscan	6BD-64 5/23-II	—
Bandpass	Texscan	6BE-50/4C-II	—
Bandpass	Texscan	6BC-64 5/23-II	—
Crystal bandpass	TMC Systems	FIL-0514	—

\* $V_{bd}$  = static breakdown voltage

<sup>13</sup>R. Sherman et al, *EMP Engineering and Design Principles*, Bell Laboratories, Electrical Protection Department (1975), 110-113.

<sup>14</sup>Capcon, Inc., *Lossyline EMI Absorptive Filters*, Cat. No. LSF-6 (no date).

<sup>15</sup>Corcom, Inc., *RFI Power Line Filters*, Cat. No. 763A.

<sup>16</sup>L. W. Ricketts et al, *EMP Radiation and Protective Techniques*, John Wiley and Sons (1976).

TABLE 2 FILTER MANUFACTURING COMPANIES

Company	Low pass	Bandpass	High pass	Band rejecting	Power line	Microwave	Shielded room
AEL Service Corp	x	x	x	x			
Allen-Bradley	x	x			x		
AVX Ceramics, Inc	x	x	x		x	x	x
Axel Electronic Inc	x	x	x	x	x	x	x
Captor Corp	x	x	x	x	x		
Erie Technological Products	x				x		
Genisco Tech	x	x		x	x		x
ITT Cannon Electric	y						
Potter Co	x				x		
RF Intertronics Div	x	x	x	x	x	x	x
San Fernando Electric	x		x		x		x
Spectrum Control, Inc	x				x		
USCC Centralab	x				x		x
Watkins-Johnson Co	x	x	x	x		x	
Airtron Div., Litton Industries	x	x	x	x		x	
Burr-Brown Research Corp	x	x	x	x			
Corcom	x		x	x	x	x	x
Corry Micronics, Inc	x	x	x			x	x
Deutsch Co	x						
ESC Electronics Corp	x	x	x	x			
General Radio	x	x				x	
Hewlett-Packard Co	x	x				x	
Hopkins Engineering Co	x	x	x				x
K & L Microwave, Inc	x	x	x	x		x	
Lorch Electronics Corp	x	x	x	x		x	
Lundy Electronics & Systems Inc	x			x	x	x	x
Maury Microwave Corp	x	x	x	x		x	
Mu-Del Electronics	x	x	x	x			
Narda Microwave Corp	x		x			x	
Osborne Electric Co., Ltd	x	x	x	x	x		
REL Industries, Inc	x	x	x	x	x		
RLC Electronics, Inc	x	x	x	x		x	
Solar Electronics Co	x	x	x	x	x		x
Sprague Electric Co	x	x	x	x	x		x
TT Electronics, Inc	x	x	x	x		x	x
Texscan Corp	x	x	x			x	
Washington Technological Assoc., Inc	x	x	x	x		x	
Rohde & Schwartz	x						
American Trans-Coil Corp	x		x			x	
EIP, Inc	x					x	
Spectral Dynamics Corp	x						
Scientific Leasing Service			x			x	
Rental Electronics, Inc			x			x	
Electro-Mechanics Co				x	x	x	
Cornell-Dubilier					x		x
Elgar Corp					x		x
Erik A. Lindgren Assoc					x		x
Polarad Electronics						x	
Solitron Devices, Inc						x	
Systron-Donner Corp						x	
Technical Research & Mfg. Co						x	
Ray Proof Corp							x
Versitron							x

TABLE 3 FILTER COST VERSUS POWER HANDLING CAPABILITY

Filter No	Current rating (A)	Cost (small quantities) (\$)
A2 <sup>a</sup>	2	28
A5	5	34
A20	20	43
A60	60	54
C25 <sup>a</sup>	25	165
C50A	50	220
D100 <sup>b</sup>	100	330

<sup>a</sup>Tested by HDL<sup>b</sup>Large attenuation at low frequencies

Care should be taken to insure that the inductor cores do not saturate, especially at the peaks of the ac power signal. Figure 6 shows that the filter cutoff frequency increases when the inductor cores saturate. Commercial filters should be tested at frequencies up to 100 MHz to insure that no resonances occur in the filter frequency response. Voltage gain does not, however, necessarily imply a power gain, since the power gain depends on the load and source impedances (eq (2)). The parasitic inductance of capacitors can also cause resonances at high frequencies. The use of lower-value, high-frequency capacitors in parallel with the regular filter capacitors can insure the proper performance of the capacitors at high frequencies (50 to 1000 MHz).

The construction and installation of a filter is as critical as its design. The input and output must be isolated, and a good filter is designed in three separate electromagnetic sections (fig 10(b)). Most filters have the ground return connected to the filter case. It is important that a good internal bond of short length be used, to avoid parasitic inductance. Also, mutual coupling between components and input and output terminals should be considered.<sup>2</sup> Because of these subtleties of filter design the purchase of commercial filters should provide better filter characteristics than a filter built from separate components. The high attenuation rate of ac filters requires multisection filters, and the large number of components adds to the difficulty of building an ac filter. The filter case as shown in figure 4 must make a tight peripheral contact to the grounded equipment cabinet to provide a nearly zero impedance. Even "hairline" apertures can greatly increase the ground impedance and change the filter characteristic.

### 3.2 Direct Current Power System

The dc power system may be considered a part of the communication equipment, in contrast to the ac power systems, which are part of the building housing the communication equipment. The dc power source for large, permanent communication facilities often consists of

<sup>2</sup>IIT Research Institute, DNA EMP Awareness Course Notes, DASA 01-69-C-0095, Defense Nuclear Agency, DNA 2772T (August 1971), 137

a large polyphase rectifier and a battery of storage cells. In newer installations this rectifier/battery system may be designed as an uninterruptible power system (UPS) whose dc output is used to supply the dc requirements of the facility and, through a dc-powered inverter, the essential ac requirements of the communication equipment

A schematic of the principal components of a dc supply system for a permanent communication facility is shown in figure 11. The supply transformer and rectifier, shown as a single-phase unit in figure 11, are usually polyphase units because of the larger conversion efficiency and smaller ripple of polyphase rectifiers. The ac power for the dc supply is obtained from the essential power bus so that the dc system can be supplied from the emergency generator in the event of a commercial ac power failure

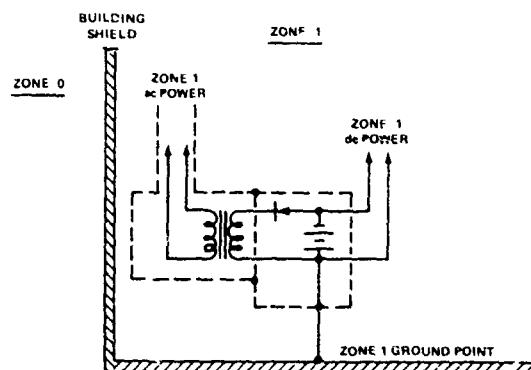


Figure 11 Schematic of dc power supply for communication facility

A typical UPS is shown in figure 12. The primary difference between the UPS and the ordinary dc system shown in figure 11 is that the UPS is designed to provide all essential power—both ac and dc—for the communications equipment. The essential controlled-environment ac power is delivered by an inverter driven by the rectifier/battery system. The raw ac power to the UPS must be obtained from an essential power bus in order for it to be supplied by either the commercial utility or an emergency engine generator. The storage battery in the UPS can usually supply the system power requirements for only a few hours without recharging

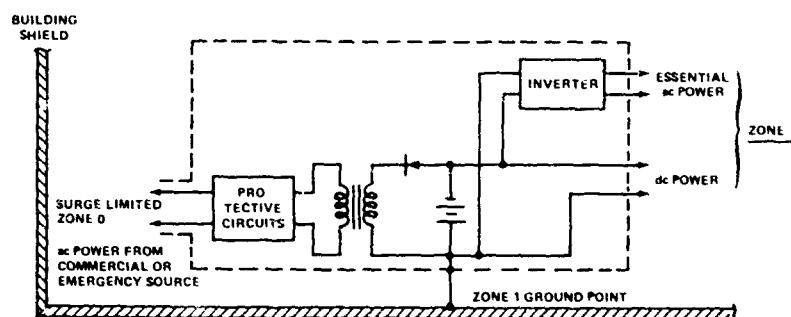


Figure 12 Uninterruptible power supply for commercial facility

<sup>1</sup>E. F. Vance et al., *Technical Inputs and EMP Design Practices for Istrasite Cabling of Telecommunication Facilities*, Stanford Research Institute, prepared for HDL under contract DAAG39-76 C-0021 (June 1977) Ch IV

The inner boundary for the dc power system is the wall of those cabinets requiring a small-signal (zone 2) environment (fig. 13). The barrier between the zone 1 dc power environment and the zone 2 small-signal region is commonly provided by line filters or dc-to-dc converters. These items are usually provided by the equipment manufacturers and may be necessary to control system-generated interference that would otherwise be distributed through the dc power conductors.

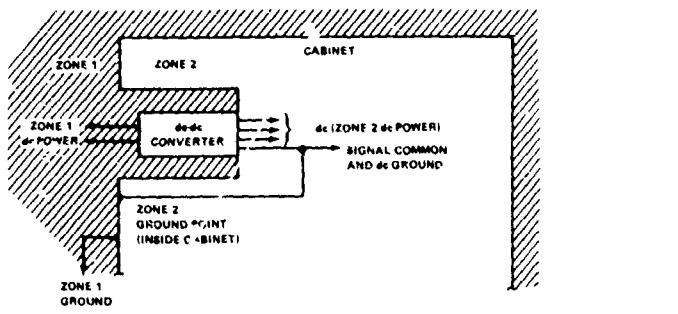


Figure 13 Power supply isolation using dc-dc converter

A common penetration of dc power conductors through a cabinet wall is illustrated in figure 13, where zone 1 dc power supplied to small-signal circuits is isolated from the small-signal region by a dc-to-dc converter.<sup>1</sup> The inner zone boundary for the dc power penetration can thus be drawn through the converter output filters or voltage regulators. Similar isolation can be provided by filters (on both conductors) at the cabinet entry point when the unconverted dc supply voltage is used by the zone 2 equipment. In a poorly shielded facility, or where large disturbance signals can be coupled to the dc supply from an ac source, filters may be needed to protect the input to the dc-dc converter. Note that the dc power lines to the dc-dc converter in figure 13 are not grounded to the equipment cabinets. When unconverted dc supply lines are used, the ground return on both filters should be referenced to the zone 2 ground point.

The dc filter component specifications are similar to the ac specifications except that capacitors need only be rated at 200 Vdc. The component specifications are repeated below.<sup>11</sup>

a. Filters should employ inductive inputs and capacitive outputs. The inductor assures firing potential for the preceding arrester and limits the current through the filter capacitor. The input inductor in these filters should be single-sweep wound and should employ adequate spacing between start and finish leads to withstand a 5000-V transient breakdown test. It should have its core taped with suitable insulating material (preferably Teflon tape) to withstand 5000 V from the winding to the core. Cores should be of powdered iron or molybdenum permalloy material. The inductor windings should employ heavy Formvar or heavy Soidereze wire only.

b. The capacitors in these filters should be of the self-healing Mylar or metallized Mylar construction and should have a breakdown voltage of not less than 200 Vdc.

<sup>1</sup>E. F. Vance et al, *Technical Inputs and EMP Design Practices for Intrasite Cabling of Telecommunication Facilities*, Stanford Research Institute, prepared for HDL under contract DAAG39-76-C-0021 (June 1977), Ch IV.

<sup>11</sup>H. A. Lasitter et al, *Nuclear Electromagnetic Pulse Protective Measures Applied to a Typical Communication Shelter*, Naval Civil Engineering Laboratory (April 1970) AD-707-696, 73 and 74.

c Potting material should be high-temperature (125 C) wax or heat-conductive epoxy  
Oil (except for flame-retardant silicon oils) or foam material should not be used

The use of low-voltage electrolytic capacitors, especially those using tantalum dielectrics, should be avoided since they are susceptible to damage from EMP-induced signals, especially under reversed polarity pulses

Because the dc power system must have sufficient capacity to supply the power requirements of equipment racks at voltages to 50 V,<sup>1</sup> very large conductors are required. Therefore, the dc distribution system may be large and bulky. Filters used with the dc power system must be capable of handling large currents because of the saturation of the magnetic cores of inductors.

One must know from coupling measurements or calculations the attenuation required to prevent damage from the EMP pulse and must assess component damage to know the effectiveness needed from a filter. Filters may normally be used on ac lines to prevent transients in normal circuit operation from propagating to other equipment. Without a detailed knowledge of the required attenuation, the use of the ac requirement of 60- to 80-dB attenuation at 100 kHz seems reasonable. However, large attenuation only at higher frequencies may be acceptable for some situations. Figure 14 shows specifications and attenuation curves from commercial filters<sup>14</sup> that can carry large currents and provide good attenuation at 100 kHz. The filters shown in figure 14 are relatively expensive because of the low cutoff frequency and high current carrying capacity. The C50 filter costs ~\$220 and the C25 filter costs = \$165 for small quantities.<sup>14</sup> Other dc filters are commercially available (see table 2)

Simple filters composed of a few elements can be effective dc filters since the cutoff frequency can be made very low. Therefore, the rapid attenuation of a multi-section filter is not required. Figure 15 shows a capacitor used as a filter in conjunction with the circuit load and source impedance. The response of this filter is

$$\frac{V_{out}}{V_{in}} = \frac{R_2/(R_1 + R_2)}{\frac{R_1 R_2}{R_1 + R_2} C j \omega + 1}$$

Since the rate of attenuation is 20-dB per decade, the cutoff frequency to provide 60-dB attenuation at 100 kHz is 100 Hz. Therefore, the capacitance value must be

$$C = \frac{R_1 + R_2}{R_1 R_2 \times 2\pi \times 100} \quad (4)$$

For 100-ohm load and source impedances,

$$C = \frac{200}{100^2 \times 2\pi} = 3.2 \times 10^{-5} = 32 \mu\text{F} \quad (5)$$

<sup>1</sup>E. F. Vance et al, *Technical Inputs and EMP Design Practices for Intrasite Cabling of Telecommunication Facilities*, Stanford Research Institute, prepared for HDL under contract DAAG39-76-C-0021 (June 1977), 133

<sup>14</sup>Capcon, Inc., *Lossyline EMI Absorptive Filters*, Cat No LSF-6

FILTER	ELECTRICAL CHARACTERISTICS							PHYSICAL CHARACTERISTICS					
	P/N	RATED CURRENT (AMPERES)	RATED VOLTAGE (VOLTS)	LINE FREQ (Hz)	MAXIMUM VOLT DROP (VOLTS)	REACTIVE CURRENT (AMPERES)	SHUNT CAPACITANCE (mF)	ATTENUATION CURVE	LENGTH (in)	WIDTH (in)	HEIGHT (in)	WEIGHT (lb)	CONFIGURATION
C25	30 25 25	600 125 125	DC 60 400		NEGLIGIBLE 0.97 3.6	0.17 0.17 1.15	4.25 4.25 4.25	A	9 7/8	2 7/8	2 13/16	5	A
C30-1	35 30 30	600 125 125	DC 60 400		0.56 0.9 2.5	NEGLIGIBLE 0.36 2.3	8.25 8.25 8.25	B	9 1/2	3 1/2	6 1/4	6.6	B
C30 AND C50	55 50 50	600 250 250	DC 60 400		0.5 1.0 6.0	NEGLIGIBLE 0.36 2.3	8.0 8.0 8.0	C	22	4	4	20	C

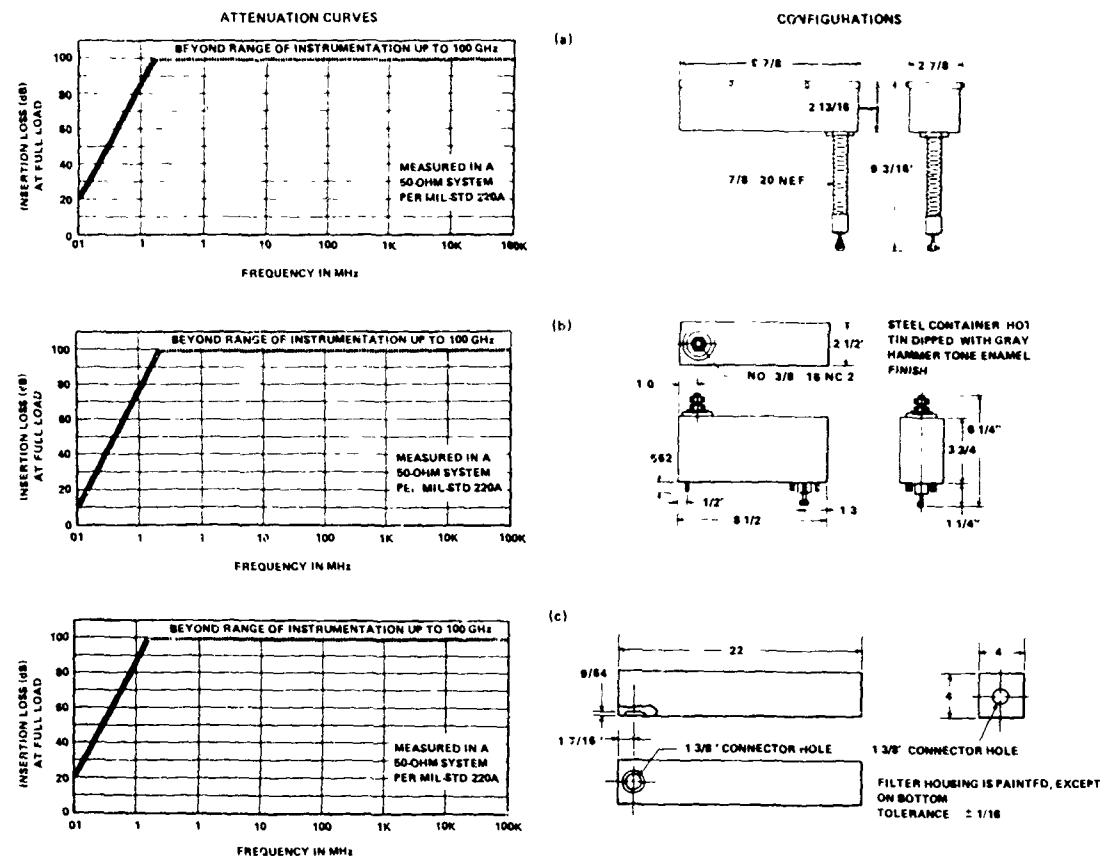


Figure 14 Specification and insertion loss of commercial filters

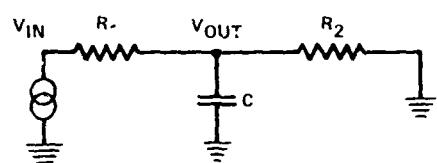


Figure 15 Simple RC filter

The value of C depends on the load and source impedance, and these impedances must be at least approximately known to use this simple formula. An electrolytic capacitor is needed to produce this large capacitance value. To reduce the effects of parasitic inductances and also voltage damage to the electrolytic capacitors, a small capacitor should be placed in parallel with the large capacitor.

Figure 16 shows a filter with a single inductor. The response of this filter is

$$\frac{V_{out}}{V_{in}} = \frac{R_2(R_1 + R_2)}{\frac{L\omega}{R_1 + R_2} + 1} \quad (6)$$

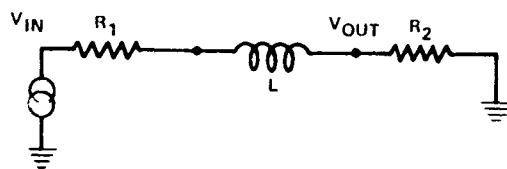


Figure 16 Simple LR filter

The attenuation rate after the cutoff frequency is again 20 dB per decade so

$$L = \frac{R_1 + R_2}{2\pi \times 100} \quad (7)$$

provides 60-dB attenuation at 100 kHz. For a 100-ohm load and source impedance

$$L = \frac{200}{200\pi} = 0.318 \text{ H.} \quad (8)$$

This is a fairly large inductance and could be expensive and bulky.

Adding a capacitor to the circuit in figure 16 produces figure 17, which is a second-order filter with an attenuation rate of 40 dB per decade. The response of this filter is

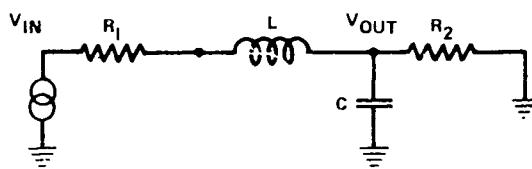


Figure 17 Simple LRC filter

$$\frac{V_{out}}{V_{in}} = \frac{R_2/(R_1 + R_2)}{\frac{R_2 LC \omega^2}{R_1 + R_2} + \left( \frac{R_1 R_2 C}{R_1 + R_2} + \frac{L}{R_1 + R_2} \right) j\omega + 1} \quad (9)$$

In order to provide 60-dB attenuation at 100 kHz with a 40-dB per decade attenuation rate, a cutoff frequency of 300 Hz is required. Therefore, the required inductance is

$$L = \frac{R_1 + R_2}{R_2 C (600\pi)^2} \quad (10)$$

For a 100-ohm load and source impedance and a 10- $\mu$ F capacitor, the inductance is

$$L = 56 \text{ mH} \quad (11)$$

Thus, a significant reduction in the inductance value can be achieved by adding a capacitor to the filter circuit. A small capacitor in parallel with the 10- $\mu$ F capacitor will reduce the effects of parasitic inductance of the large capacitor.

A simple means of providing inductance is the use of ferrite shield beads (sect. 3.3.2). These beads present several problems in dc filters. First the small shield beads produce low inductance values (about 1  $\mu$ H) for a single turn, and they are subject to saturation from dc currents (fig. 18). A new material called EMI suppressant tubing<sup>14</sup> has a high dc saturation and provides shielding as well as filtering. However, it is most effective (compared to beads) above 100 MHz, which is above the frequency range where most of the EMP energy is concentrated (fig. 19). Long lengths of this tubing can be used which can provide better low-frequency attenuation and can also provide shielding to prevent direct pickup in poorly shielded facilities.

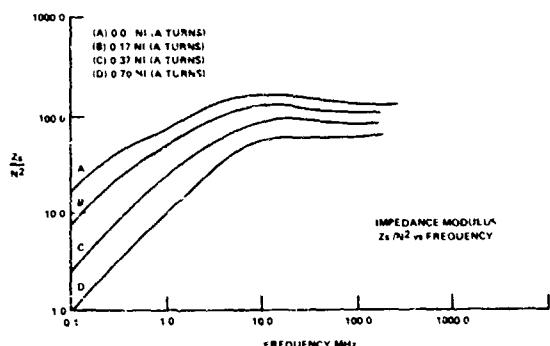


Figure 18 Impedance of ferrite bead with different dc currents

<sup>14</sup>Capcon, Inc., Lossyline EMI Absorptive Filters, Cat. No. LSF-6

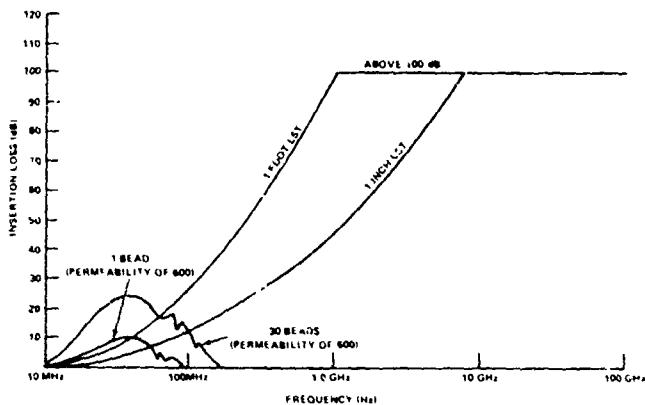


Figure 19 Insertion loss of EMI suppressant tubing and ferrite beads

Figure 20(a) shows a bifilar and an oppositely wound choke which are effective at low frequencies and can prevent dc current saturation with a balanced pair of wires,<sup>13</sup> since the magnetic flux from one winding cancels the flux from the other winding. Because of the flux cancellation of differential signals, these chokes are primarily effective for common-mode signals. Figure 20 (b) shows a filter for a dc power line.<sup>3</sup> The coaxial capacitors and capacitors C1 and C2 short out both differential and common-mode ac signals. The input and output leads should be separated to prevent arcing. A reasonably large inductance value can be achieved with only a relatively few turns because of the high permeability of the ferrite core. The number of windings on a toroidal core to produce a given inductance can be found for manufacturers' catalog ratings using the following equations.<sup>6</sup>

$$N = \sqrt{\frac{L}{L_0}} \times 10^3, \quad (12)$$

where

$N$  = number of turns

$L_0$  = inductance for 100-turn winding, and

$L$  = required inductance

<sup>3</sup>R. A. Perala and T. F. Ezeil, *Engineering Design Guidelines for EMP Hardening of Naval Missiles and Airplanes*, Mission Research Corp., prepared for the Naval Ordnance Laboratories under contract N60921-73-C-0333 (December 1973), Ch 7, p. 11.

<sup>6</sup>Philip R. Geffe, *Simplified Modern Filter Design*, Hayden Book Company, Inc. (1963), 100.

<sup>13</sup>R. Sherman et al, *EMP Engineering and Design Principles*, Bell Laboratories, Electrical Protection Department (1975), 107.

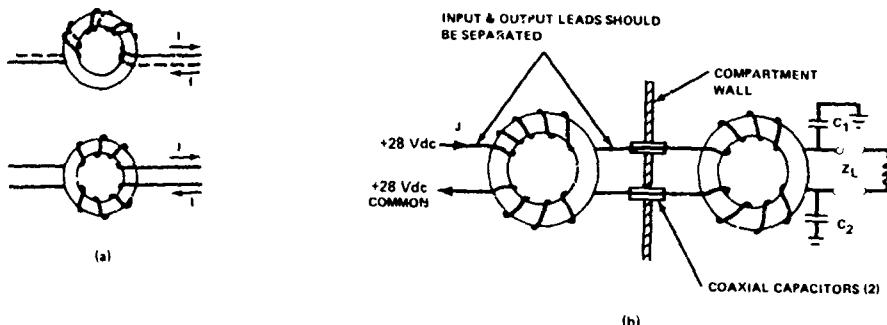


Figure 20 Simple filters for common-mode suppression (a) bifilar (top) and standard (bottom) wound transformers and (b) nonsaturating dc power-line filter

As an example, a particular ferrite core with a high permeability has a value of  $L_o$  of approximately 550  $\mu$  mH per 1000 turns.<sup>17</sup> Therefore, a 1-mH inductance requires

$$N = \sqrt{\frac{1}{5500}} \times 10^3 = 14 \text{ turns} . \quad (13)$$

Another consideration is the saturation of the core. Calculating the saturation of the core requires knowing the maximum current through the coil, which in turn requires the applied signal and the value of the other components in figure 20. The use of circuit analysis codes greatly simplifies the calculation of the current.<sup>7,8,9</sup> The maximum B field for this core is given as 3000 Gauss. However, this value is a function of temperature. The value of the B field can be calculated from the following equation

$$B_{max} = \frac{LI_{max}}{NA_e} \times 10^8 , \quad (14)$$

where  $L$  = core inductance, H, and  $A_e$  = effective area of core,  $\text{cm}^2$

The value of  $A_e$  can also be found in inductor core catalogs.<sup>17</sup> So if the value of  $B_{max}$  calculated is greater than 3000 Gauss, the core will saturate, and a different core material will have to be tried. The modeling of the winding as a pure inductance is an approximation since the ferrite core is lossy and frequency dependent. Note also that in figure 20(b) two filters are used, one on each of the balanced pair of wires. Each filter is referenced to ground through the capacitors. Because of the mutual coupling of the cores, which prevents dc saturation, the inductance of the filters is ineffective for differential signals. The capacitors are, however, still effective for filtering differential signals.

<sup>7</sup>Allan F. Malmberg, NET-2 Network Analysis Program Release 9, BDM Corporation, prepared for HDL under contract DAAG39-70-C-0050, HDL-050-1 (September 1973)

<sup>8</sup>Laurence Nagel, SPICE-2: A Computer Program to Simulate Semiconductor Circuits, Electron Research Laboratory, College of Engineering, University of California at Berkeley (May 1976)

<sup>9</sup>L. D. Millman et al, CIRCUS: A Digital Analysis Program of Electronic Circuits—Program Manual, The Boeing Corporation, prepared for HDL under contract DA-49-186-AMC-346(X), Harry Diamond Laboratories, HDL-346-2 (January 1967)

<sup>17</sup>Ferroxube Corporation of America, Applying Ferroxube Ferrite Cores to the Design of Power Magnetics Bulletin 330 (1966)

### 3.3 Analog Signals

The entrance of signal cables with voice or modulated carrier signals into a facility requires large bundles of wires. These wire bundles can be coupled to disturbance signals either directly or through penetrators which breach the facility housing. The signal pickup on the wire bundles is often a common-mode signal, and the use of common-mode rejection techniques such as balanced lines, isolation transformers, and series transformers can be used to reduce these disturbance signals. HEMP-coupled transients can also be reduced by the following techniques (1) shielding the wire bundle, (2) decoupling the small-signal conductors from the HEMP environment by using twisted pairs, (3) separating small-signal cable from other intrasite cables (grouping), and (4) disorienting small-signal cables (orthogonalizing). These techniques are illustrated in figure 21.<sup>1</sup> The use of shielded cable affords the most complete protection, but it is usually the most expensive method. The use of twisted pairs is effective in preventing differential- or signal-mode coupling, but common-mode interference induced on the pair must be accommodated (rejected) at the terminal circuits. Circuit separation or grouping in the cable trays performs the same function (isolation) as a cable shield, but it is generally much less effective than a shield. Grouping, in which cable groups within a tray are separated by metal septa, more nearly approaches the effectiveness of a cable shield, but with a commensurate increase in cost. Orthogonalizing small-signal cables (that is, routing them perpendicular to other cables) can, in principle, be moderately effective, but in practice there are many instances where it cannot be applied.

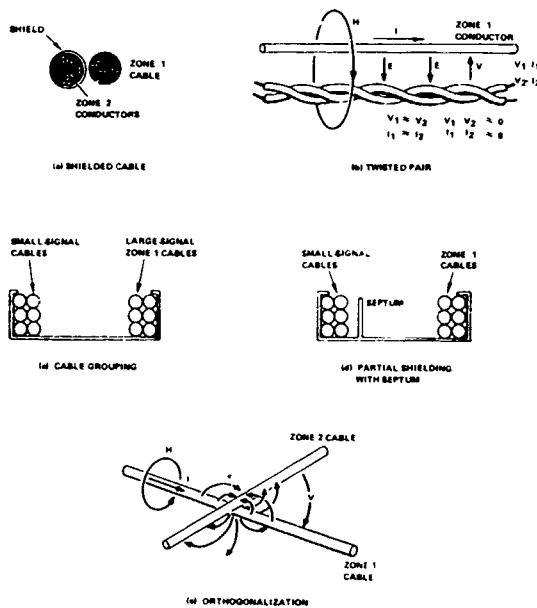


Figure 21 Reducing pickup on small-signal cable lines

<sup>1</sup>E. F. Vance et al, *Technical Inputs and EMP Design Practices for Intrasite Cabling of Telecommunication Facilities*, Stanford Research Institute, prepared for HDL under contract DAAG39-76-C-0021 (June 1977) Ch VII

The hardening of ac and dc power lines using filters depended on an attenuation guideline of 60 to 80 dB at 100 kHz. Such a guideline would probably be effective for analog signal lines, but may in many instances be an overkill. The analog signal-line information will usually have a higher frequency content than ac power frequencies, so a low-pass filter with a low cutoff frequency can affect the signal information. Bandpass filters and tuned receiver amplifiers are less likely to interfere with normal signals, provided that the normal signals occupy a relatively narrow frequency band. The use of simple filter elements, especially small ferrite beads which create a lossy inductor when placed on wires, is a relatively simple and inexpensive means of circuit hardening. Care must be taken to determine whether the hardening is sufficient and does not interfere with normal circuit operation. This will require a more detailed analysis of EMP coupling to the wires, circuit operation, and component damage levels.

### 3.3.1 Cable Shield Coupling Reduction with Magnetic Cores

Large cable shield currents can couple to the analog signal wires. Placing cores around the shields has the following benefits.<sup>11</sup> The magnetic cores should

- (a) be easy to install on existing cables,
- (b) introduce a maximum of inductance into the cable,
- (c) not saturate under the highest cable current,
- (d) be of such nature that flux is not trapped in the core, and
- (e) introduce electrical losses for any oscillatory pulse current flowing on the cable

Benefit (a) implies that cores must be split into two parts or otherwise be capable of disassembly.

Benefit (b) implies that each ampere of current through the center of the core must set up a maximum amount of magnetic flux since  $L = N(d\phi/dI)$ . For a given core size this means that the core permeability should be a maximum and the reluctance of the magnetic circuit be as low as possible. These, in turn, imply that the mean circumference of the core must be as small as possible and the air gaps must be held to a minimum. For rapidly changing pulse conditions, high permeability implies that the core laminations should be thin.

These restrictions tend to be at variance with requirements (c), (d), and (e). Minimum air gaps mean that a given cable current carries the core further into saturation since a small air gap introduces some reluctance into the circuit and allows more current to flow before the core saturates. It also allows the core flux to return more nearly to zero when the current goes to zero. Thinner laminations have higher permeabilities for short duration pulses, but they cause less hysteresis loss per cycle. In this particular application, high hysteresis loss is desirable. Cores with thin laminations also cost much more than cores with standard 12-mil laminations.

An example of the use of cores will be given. A core that would fit over a communication cable has a 3.7 cm ID and an 11.7 cm OD. A  $B_{max}$  value of 10 kGauss was determined at 12 Oersted (H). The value of  $L_c$  (inductance for 1000 turns) was not given by the manufacturer, but the coil inductance can be calculated by the equation.<sup>12</sup>

<sup>11</sup>H. A. Lasitter et al. *Nuclear Electromagnetic Pulse Protective Measures Applied to a Typical Communication Shelter*. Naval Civil Engineering Laboratory (April 1970) AD-707-696, 99

<sup>12</sup>Ferroxcube Corporation of America. *Applying Ferroxcube Ferrite Cores to the Design of Power Magnetics*. Bulletin 330 (1966)

$$L = 0.4 \pi N \mu_e \left( \frac{A_e \text{ cm}}{l_e \text{ cm}} \right) \times 10^{-8} \text{ Henries} , \quad (15)$$

where  $\mu_e$  = effective permeability,  
 $A_e$  = effective area of core  $\text{cm}^2$ , and  
 $l_e$  = length of flux path, cm.

Since we want to find the saturation current, the average value of  $\mu_e$  is used, which is the straight-line slope between  $B_{max}$  and  $H_{max}$ . The value of L for one turn, assuming the cable passes through the choke, is

$$L = 0.4\pi \times \frac{10^4}{12} \left( \frac{12.9}{29} \right) \times 10^{-8} = 4.6 \mu\text{H} , \quad (16)$$

and from equation 14,

$$I_{max} = \frac{NB_{max}A_e}{L} \times 10^{-8} = \left( \frac{10^4 \times 12.9}{4.6 \times 10^{-8}} \right) \times 10^{-8} = 2.8 \times 10^3 \text{ A} \quad (17)$$

Figure 22 shows a series of chokes on a cable with various grounds. The attenuation of the chokes depends on the inductance and resistance of the ground leads as well as the inductance of the chokes.<sup>11</sup>

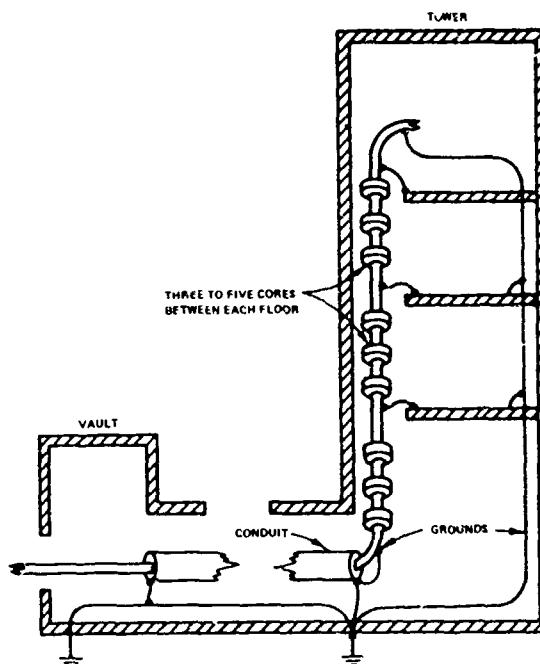


Figure 22 Communication cable using choke cores and multiple grounds to eliminate shield currents

<sup>11</sup>H. A. Lasitter et al, Nuclear Electromagnetic Pulse Protective Measures Applied to a Typical Communication Shelter, Naval Civil Engineering Laboratory (April 1970) AD-707-696, 146

### 3.3.2 Ferrite Shield Beads

Ferrite shield beads, when placed on a wire, have the equivalent circuit of an inductor and resistor (fig. 23). They are commercially available as small toroidal cores of various sizes (fig. 24). The beads have the advantages of being easy to fit over existing wires, of low cost, and of being dissipative rather than reflective. Ferrite cores for attenuation of high-frequency signals are available from many companies.<sup>18-20</sup> They have the disadvantage of current saturation either from a disturbance signal or from normal signal levels (fig. 18). Also, the resistance and inductance is frequency dependent (fig. 25), and every core material will have a different characteristic. This frequency dependence can be difficult to model analytically, but it is useful for increasing dissipative losses at high frequencies. The information for manufacturers to describe the characteristics of a particular bead can vary greatly. Often, the total impedance of the bead on a wire is given as a function of frequency (fig. 25). This total impedance can be used to estimate the attenuation when using a bead or to determine the number of beads required for a given attenuation. The insertion loss can be estimated using figure 26.

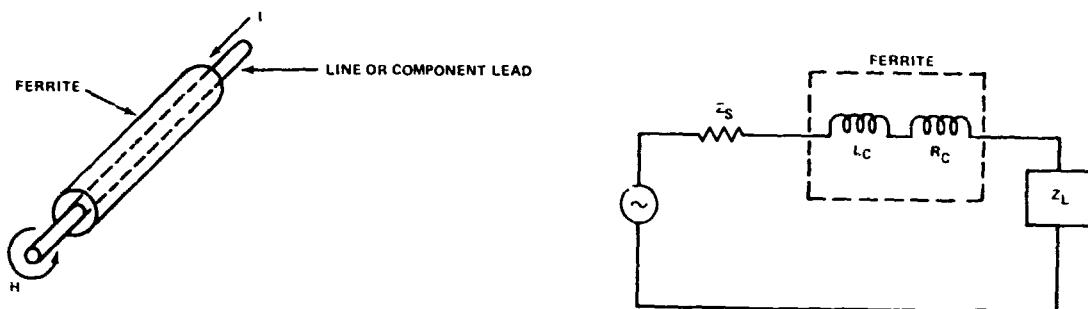


Figure 23 Ferrite bead on wire and ferrite bead equivalent circuit

DIMENSIONS (INCHES)		
O D	I D	HEIGHT
0.210	0.052	0.160
0.210	0.052	0.400
0.260	0.080	0.160
0.260	0.080	0.400
0.200	0.100	0.160
0.200	0.100	0.400
0.395	0.105	0.160
0.395	0.105	0.400
0.085	0.020	0.0625

DIMENSIONS (INCHES)		
O D	I D	HEIGHT
0.200	0.050	0.160
0.200	0.050	0.400
0.250	0.075	0.160
0.250	0.075	0.400
0.200	0.100	0.160
0.200	0.100	0.400
0.375	0.100	0.160
0.375	0.100	0.400
0.062	0.017	0.0625

Figure 24 Commonly available sizes of ferrite beads

<sup>18</sup>Ferronics, Inc., Wide Band Ferrite Cores, Bulletin 401, New York (no date)

<sup>19</sup>Steward Manufacturing Company, Technical Bulletin Ferrite Shielding Beads, Chattanooga, TN (no date)

<sup>20</sup>Indiana General, Ferrites Catalog 209, Keasbey, New Jersey (no date)

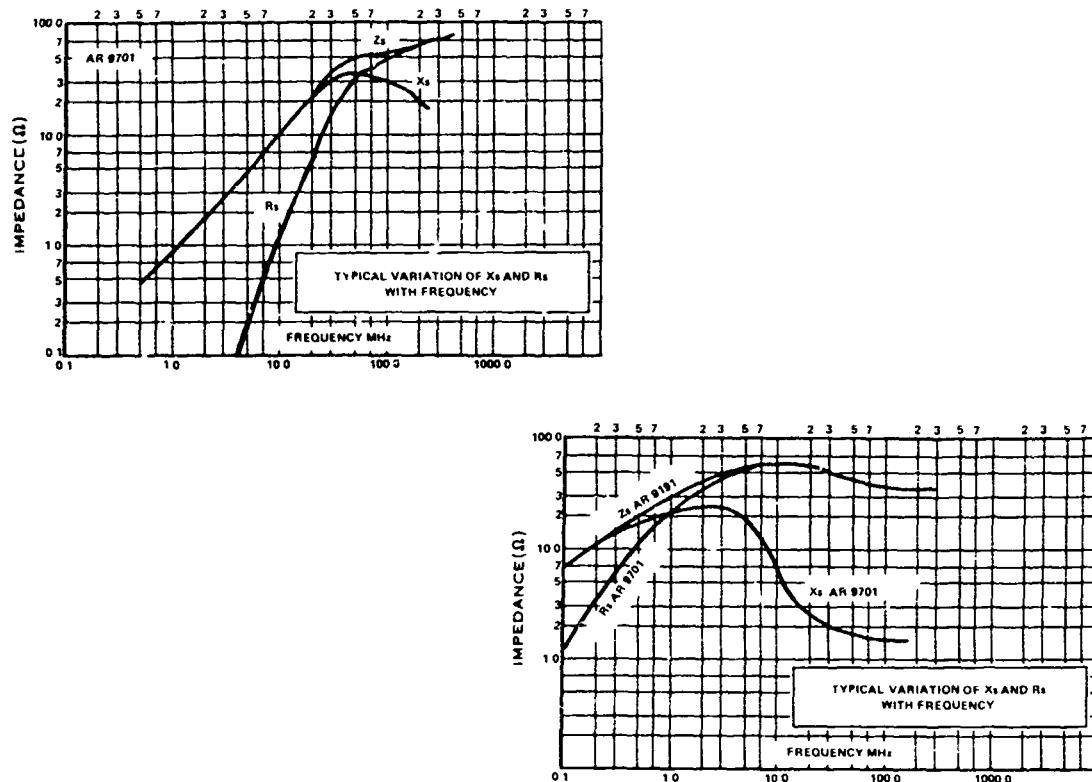


Figure 25 Ferrite bead frequency-dependent resistance and reactance for two ferrite materials

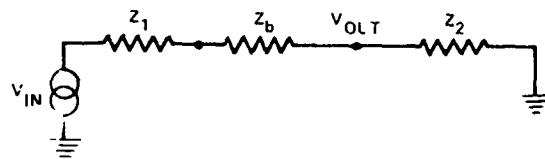


Figure 26 Circuit for estimating insertion loss

$$\text{Attenuation loss} = 20 \log \frac{V_{\text{out}}}{V_{\text{in}}} = 20 \log \left| \frac{Z_2}{Z_1 + Z_b + Z_2} \right| \quad (18)$$

where  $Z_b = \sqrt{R^2 + X^2}$ .

The larger  $Z_b$  is relative to  $Z_1$  and  $Z_2$ , the greater the attenuation. If a more exact frequency response, including possible resonances between the reactances of  $Z_1$ ,  $Z_2$ , and  $Z_b$  is desired, the NET-2<sup>7</sup> circuit-analysis code can be used. NET-2 has a tabulated transfer function which can be used to model the bead impedance in the frequency domain.<sup>10</sup> In a circuit, the bead impedance

<sup>7</sup>Allan F. Malmberg, NET-2 Network Analysis Program Release 9, BDM Corporation, prepared for HDL under contract DAAG39-79 C-0050, HDL-050-1 (September 1973)

<sup>10</sup>BDM Corporation, NET-2 Network Program Analysis Addendum to User's Manual for Release 9.1, prepared for HDL under contract DAAG39-77-C-0150, BDM/W-77-573-TR (November 1977)

determines the current through the bead for a given voltage drop across the bead. The transfer function provides an output voltage equal to the voltage drop across the bead when an input voltage numerically equal to the current through the bead is the input. The circuit in figure 27 can be used to model the bead impedance in a circuit. A voltage equal to the current through the bead is produced by the THRU CURRENT element. Applying this voltage to the tabulated transfer function input provides the voltage drop across the element. Finally, this voltage drop is created in the circuit by the voltage-controlled voltage source

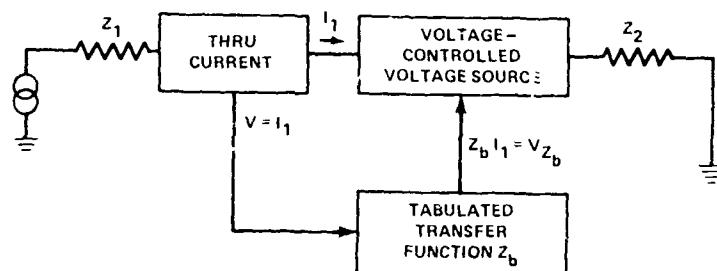


Figure 27 Equivalent circuit used with NET-2 to simulate ferrite bead in an electronic circuit

Figure 28 shows the measured inductance and resistance of a single ferrite bead on a wire. Note that the inductance decreases and the resistance increases with frequency. The net effect of the inductance decrease and resistance increase is a leveling off of the impedance at high frequencies. This impedance leveling can also be seen in figure 25. Because of the relatively low inductance and resistance values of this bead (fig. 28), the value of  $Z_1$  and  $Z_2$  was chosen as 1 ohm in the circuit used to calculate the bead response (fig. 26). The low resistance value reduces the cutoff frequency of the filter produced by the bead in the circuit. To produce a reasonable disturbance attenuation in the frequency range of interest for EMP disturbances with larger source and load impedance (50 to 100 ohms), more beads or beads with larger impedance values are required. However, because of the relatively low impedance of a ferrite bead, the beads are more effective in low impedance circuits, as is demonstrated by this example. Figure 29 shows the response of the bead used in the circuit shown in figure 25, with  $Z_1$  and  $Z_2$  equal to 1 ohm. Also shown is the response using just an inductor with the initial inductance value of the bead (300 nH). Above 4 MHz, the inductor produces greater attenuation than the simulated bead response. The beads do have the value of being dissipative rather than reflective, and more beads can easily be added to provide the attenuation required. In general, the individual impedances of beads can be added to determine approximately the total attenuation loss of the beads in a circuit. However, when using large numbers of beads, the attenuation at some frequencies may be reduced by the additional beads.<sup>3</sup> Figure 30 shows that, above 25 MHz, using 30 beads produces more attenuation than using 300 beads.

Another problem with beads is the saturation of the magnetic material. If magnetic hysteresis curves or data on the magnetic characteristics of the core are given, saturation currents can be calculated using equations 15 to 17. The effect of dc currents can be found from manufacturers' data such as the curves shown in figure 18

<sup>3</sup>R. A. Perala and T. F. Ezell, *Engineering Design Guidelines for EMP Hardening of Naval Missiles and Airplanes*, Mission Research Corp., prepared for the Naval Ordnance Laboratories under contract N60921-73-C-0033 (December 1973) Ch 7, p 10

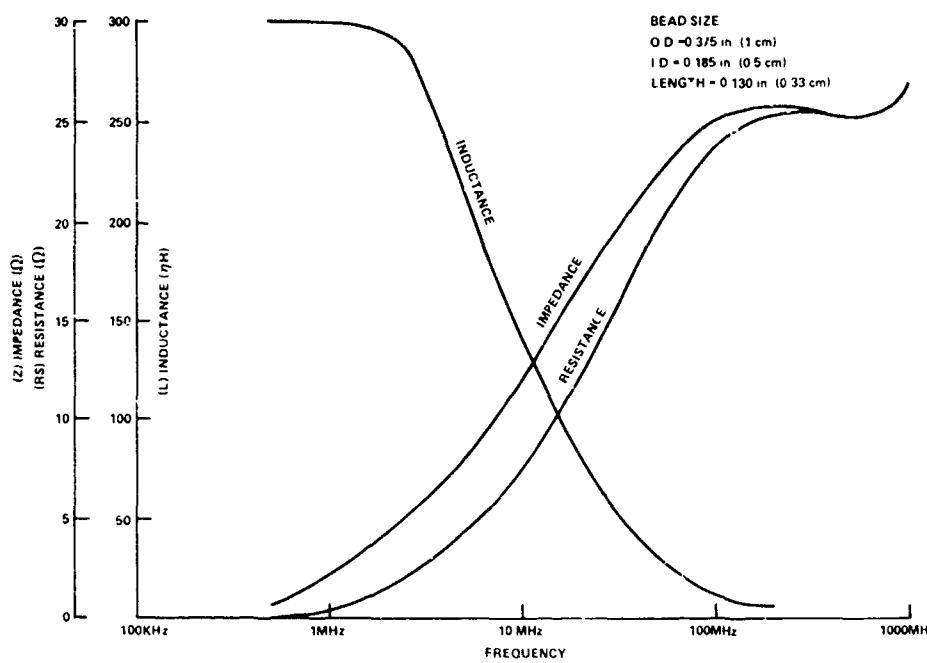


Figure 28 Measured inductance and resistance of ferrite bead

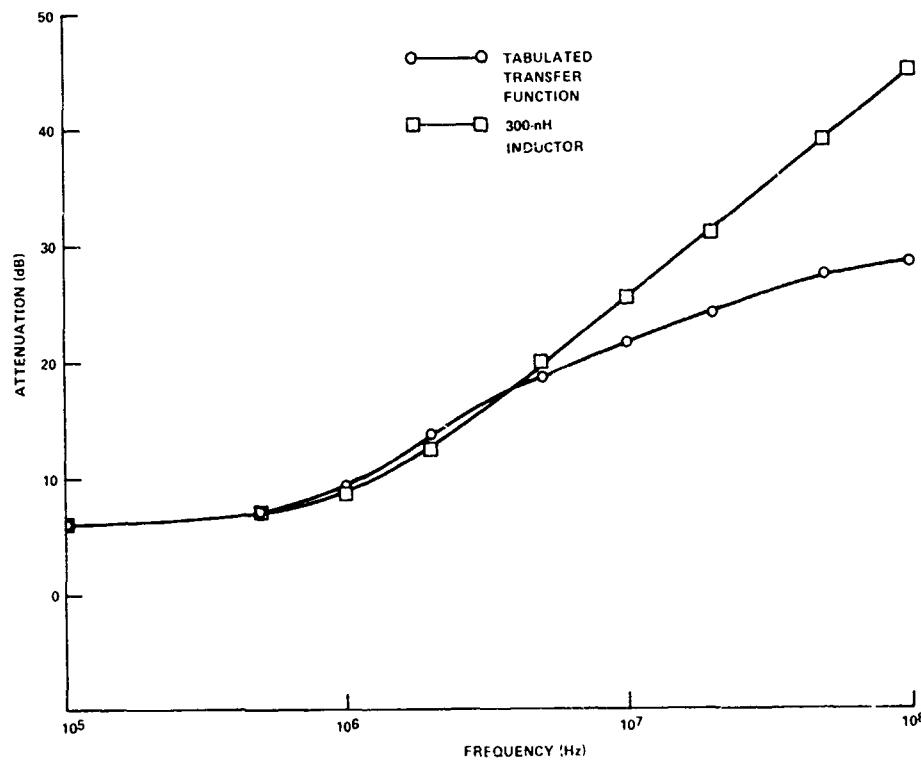


Figure 29 Response of ferrite bead simulated with NET-2 circuit-analysis program compared to low-frequency inductance model of bead

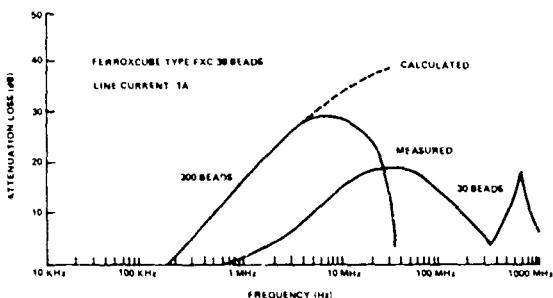


Figure 30 Attenuation with large numbers of ferrite beads

### 3.3.3 Miniature Pin Filters

Ferrite materials are also used to manufacture miniature filter pin connectors. These connectors are available in different shapes (circular or rectangular) and sizes with up to 61 contacts. The filtering of many signal lines is performed by these filters in a much smaller volume than would be required by using ordinary filters. Figure 31 shows the attenuation of some miniature rectangular filters.<sup>21</sup> The low-frequency filter (fig. 31) costs approximately \$85 in small quantities and is substantially more expensive than the other filters, which cost approximately \$45 in small quantities. Since little information is available on the EMP hardness of these filters, the testing of filters would be recommended. Of the filters whose frequency characteristics are shown in figure 31, two types have been pulse tested.<sup>22</sup> A  $\pi$ -section filter with a cutoff frequency of 2 MHz and a low-frequency filter with a cutoff frequency of 100 kHz were tested. Using rectangular pulses with widths of 100 ns and 1  $\mu$ s and amplitudes to 5 kV, no filter performance degradation was observed until arcing occurred from the pins to the case of the filter. The arcing occurred at the voltages shown in table 4.

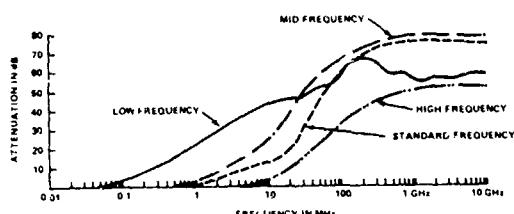


Figure 31 Attenuation of miniature pin filters

<sup>21</sup>R. A. Perala and T. F. Ezell, *Engineering Design Guidelines for EMP Hardening of Naval Missiles and Airplanes*, Mission Research Corp., prepared for the Naval Ordnance Laboratories under contract N60921-73-C-0033 (December 1973), Ch 7, p 2.

<sup>22</sup>Cannot ITT, *Miniature Rectangular Filter Contact Connectors, DPXJ/DPX\*J* (January 1973).

TABLE 4 OCCURRENCE OF ARCING AT VARIOUS VOLTAGES

Filter	100 ns	1 $\mu$ s
DEJ-9TP ( $\pi$ section)	2 kV	11 kV
DEJ-9LP (low frequency)	> 5 kV	12 kV

Unless the arcing causes permanent damage (which was not reported), arcing can actually be beneficial, since the pulse energy does not pass through the filter. In determining whether the filter characteristics of a pin connector filter are adequate, the load and source impedance for which specifications are given should be known. It is also necessary to know the load and source impedance the filter will see when installed in a signal line. If the test and actual use load and source impedances are different, testing or computer circuit simulation is necessary in order to determine the actual attenuation of the filter.

### 3.3.4 Ferrite Beads and Nonlinear TPDs

An application of ferrite beads in a filter will be shown in order to demonstrate the use of ferrite beads with nonlinear TPDs. The filter will be relatively simple, but the use of a simple filter, which uses a ferrite bead, and a nonlinear TPD (Zener diode, spark gap, etc.) can be quite effective. The electronic TPD is used to clamp the input voltage at a level slightly above the maximum normal signal level. Thus, the normal circuit operation is not affected. Since many nonlinear TPDs do not have a fast clamping response, the initial and possibly the final part of an EMP-induced input passes through the nonlinear TPD (fig. 32).<sup>3</sup> The signals that pass through the TPD are of short duration and thus, high frequency. If the pulse amplitude is sufficiently high, component damage can still occur even though the pulse is very narrow. Because of the high-frequency content of the signal that passes through, a high-frequency cutoff filter that is small and uses few components can be effective. Figure 33 shows the peak output voltages produced by input currents generated by pulses of 50 to 500 V, 4-ns risetime, and 500-ns duration, with a 25-ohm source impedance applied to the protection circuit shown in the figure.<sup>22</sup> The Zener diode TPD followed by the filter (ferrite bead and 970-pF capacitor) produced less output voltage than either the diode alone or the diode preceded by the filter. By placing the filter after the TPD, pulses passing through the TPD are then attenuated. Also, having the filter inductor before the capacitor increases the rate of voltage rise across the TPD, causing it to fire sooner.

The energy reduction produced by the TPD ferrite bead filter combination shown in figure 33 was calculated for a 50- and 1000-ohm load termination.<sup>22</sup> Using a 500-V, 500-ns wide pulse the energies are as follows.

<sup>3</sup>R. A. Peala and T. F. Ezell, *Engineering Design Guidelines for EMP Hardening of Naval Missiles and Airplanes*, Mission Research Corp., prepared for the Naval Ordnance Laboratories under contract N60921-73-C-0033 (December 1973), Ch 7, p. 10.

<sup>22</sup>R. C. Keyser, *Satellite Hardening Techniques Evaluation and Test Definition*, IRI Corp., Air Force Weapons Laboratories, AFWL-TR-74-328 (September 1975) AD-B007040L

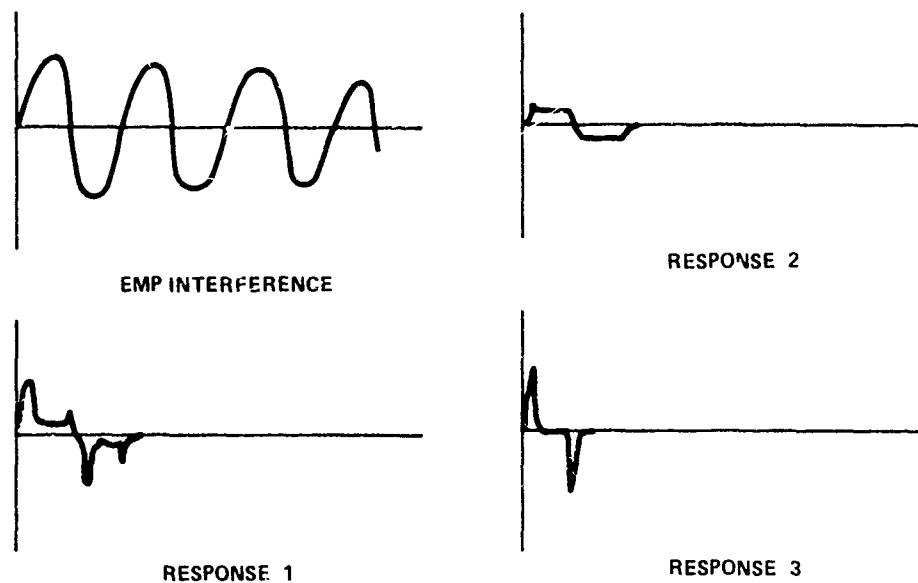


Figure 32 EMP-induced sinusoidal signal at top passes through three different TPDs

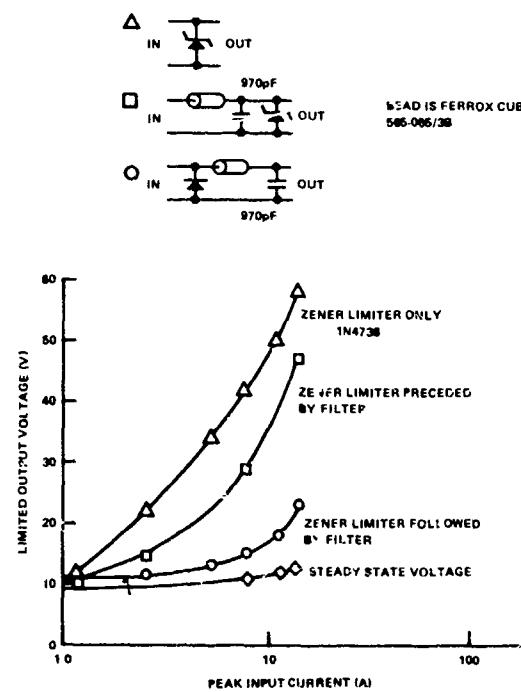


Figure 33 Peak output voltage for a TPD and TPD filter combination

TABLE 5 ENERGY REDUCTION PRODUCED BY TPD FERRITE BEAD FILTER COMBINATION

Load impedance (ohms)	Incident pulse energy (J)	Limited pulse energy (J)
50	2500	1.8
1000	130	0.09

The energy attenuation produced by the TPD ferrite bead filter is approximately 1400 (31 dB), which is a significant energy reduction. The high-frequency characteristic of the pulses that pass through the TPD requires that the parasitic inductances of filter element be small. The use of commercial feedthrough and pin connector filters previously described is most effective, since they are designed to minimize critical lead length inductance. The effect of inductance due to component line length is seen in figure 34.<sup>22</sup> The measured transfer response of the filter in figure 33 is shown with a 2.5-cm and a 0.5-cm capacitor lead length. Only the high-frequency response is affected by the longer lead length, but this can be significant in attenuating the high-frequency pulses that pass through the TPD.

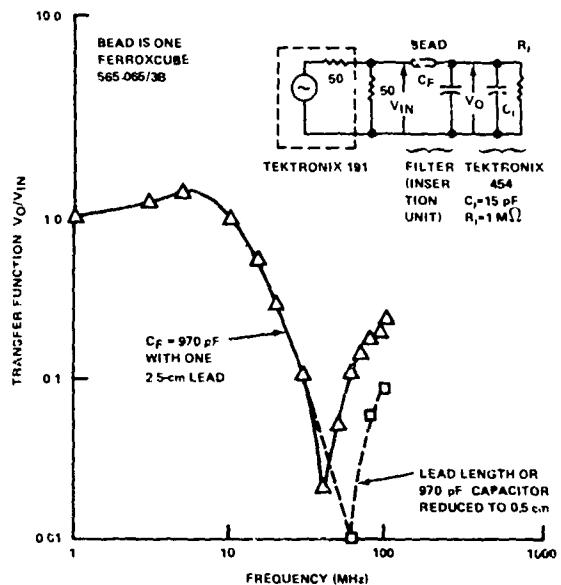


Figure 34 Effect of capacitor ( $C_F$ ) lead length on filter performance.

### 3.3.5 Balanced Analog Signal Transmission

Analog signal transmission between equipment racks can be protected from EMP-induced interference by the use of balanced lines with isolation transformers. Figure 35 shows an example of circuits that provide balanced isolation. A differential signal applied at the primary of one transformer produced a differential signal at the secondary of the other transformer. If an EMP signal produces a common-mode signal on the balanced pair, no differential output signal is

<sup>22</sup>R. C. Keyser, *Satellite Hardening Techniques Evaluation and Test Definition*, IRI Corp., Air Force Weapons Laboratories, AFWL-TR-74-328 (September 1975) AD-b007040L

produced at the transformer primaries if the transformer center tap had identical circuits to each of the balanced wires and no capacitive coupling exists between the primary and secondary windings. In practice, 30 to 40 dB of common-mode rejection can be obtained.<sup>2</sup>

The isolation from EMP-induced disturbances can be further increased by using a modulated carrier to transmit analog information. Transformers can be tuned to the narrow bandwidth of the carrier frequency, thereby reducing the broad spectrum energy transmission from the disturbance signal. By using an equivalent circuit model of a transformer, a "tuned" transformer which is effectively a bandpass filter can be designed. Figure 36 shows a transformer and the low-frequency equivalent circuit of the transformer. The resistance of the primary wires ( $R_p$ ), secondary wires ( $R_s$ ), and load ( $R_L$ ) are shown. Also shown is an equivalent core loss resistance  $R_c$  which accounts for power dissipated in the transformer core. Figures 37 (a) and (b) show a capacitor placed across the transformer primary (primary tuning) and across the transformer secondary (secondary tuning), respectively. The resonance frequency for the two types of tuning is

$$\omega_p = \frac{1}{L_p C_p} \quad (19)$$

and

$$\omega_s = \frac{n}{L_p C_s} \quad (20)$$

The Q for various cores at different frequencies is given in transformer core catalogues.<sup>23</sup> From these equations, and using a core with a reasonably good Q, a tuned transformer can be designed. For high-frequency carrier frequencies, the low-frequency equivalent circuit must be modified (fig. 36) to account for additional parasitic effects such as leakage inductance and the equivalent shunt capacitance of the interwinding capacitance.<sup>19</sup>

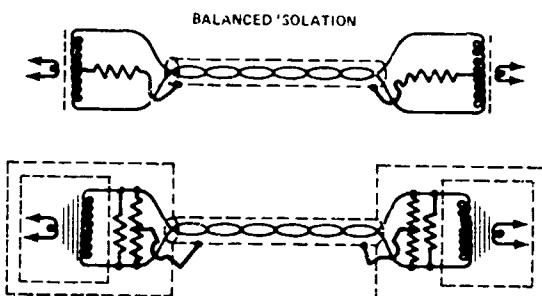


Figure 35 Balanced line isolation using transformer

<sup>21</sup>It Research Institute, DAN EMP Awareness Course Notes, DASA 01-69-C-0095, Defense Nuclear Agency, DNA 2772 (August 1971)

<sup>18</sup>Ferronics, Inc., Wide Band Ferrite Cores, Bulletin 401, New York (no date)

<sup>23</sup>Ferroxcube Corporation of America, How to Design Optimum Inductors, Bulletin 220-B (no date)

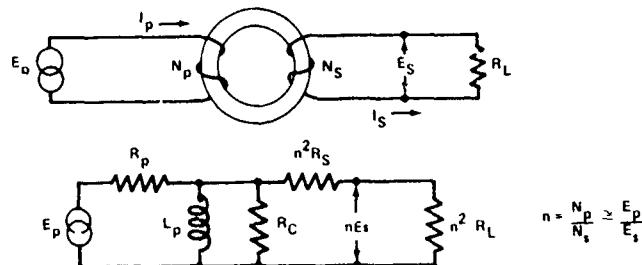


Figure 36 Transformer and equivalent circuit of transformer

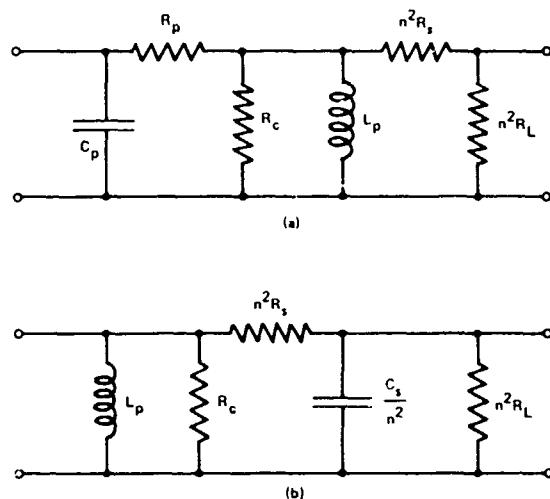


Figure 37 (a) Primary tuned transformer and (b) secondary tuned transformer

### 3.3.6 Crystal, Ceramic, and Mechanical Filters

Crystal, ceramic, and mechanical filters are basically high-Q, bandpass filters with  $b_c$  widths from 0.001 to 0.7 percent of their center frequency. The crystal and ceramic filters have spurious resonances which can allow the passage of considerable energy from a broad spectrum EMP-induced signal. Little information is available on the EMP hardening provided by these types of filters. The arcing-over of these filters when exposed to a high-voltage EMP induced signal is also possible. A crystal bandpass filter has been tested, however, at voltages to 11 kV without failure.<sup>13</sup>

<sup>13</sup>R. Sherman et al, *EMP Engineering and Design Principles*, Bell Laboratories, Electrical Protection Department (1975), 110-113

### 3.3.7 Problems with Analog Filters

A general problem with using filters to protect analog circuits from EMP damage is knowing how much attenuation is required for a given application. Unless the disturbance signal level and frequency characteristics as well as the circuit damage threshold are known, the required amount of attenuation cannot be determined. If filters are designed with large safety margins, then low cutoff frequencies, which require large-value components, or rapid rates of attenuation, which require many components, are needed. Large-value components are bulky, expensive, and likely to have parasitic reactances. The use of many components becomes expensive and bulky.

When purchasing commercial filters or designing filters from discrete components, it is necessary to know the source and load impedance for the application. If the source or load impedance has reactive impedances, then spurious resonances can occur. The wide-band energy of an EMP-induced signal can pass through the filter at the frequency of these spurious resonances, possibly causing circuit damage. The equipment used to measure filter attenuation according to MIL-STD-220A can be modified to perform scattering parameter measurements of a filter.<sup>16</sup> The scattering parameter measurements can be used to calculate the filter attenuation with any known loads.

The construction and installation of a filter is as important as the filter design.<sup>16</sup> It is essential that the filter input and output be isolated from each other. Since most filters operate relative to ground, it is important that the return side of the filter is well bonded to the filter case, which is in turn connected to a "good ground" with low-inductance straps and large connection contact areas. Arcing problems with filters and isolation transformers can also be a problem. Filters are usually rated for hundreds of volts, but an EMP-induced signal may be many thousands of volts. Table 1 lists filters that passed an 11-kV, 50-ns pulse test performed at the Harry Diamond Laboratories.<sup>2</sup> Unless these particular filters are used, it would be good practice to have an EMP filter tested with high-voltage short-duration pulses.

The reduction of common-mode voltage pickup using isolation transformers is dependent on the symmetry of the terminations. If arcing occurs at the transformer terminations, a large differential signal can be formed. Capacitive leakage between primary and secondary transformer windings can also reduce the isolation of the transformer.<sup>16</sup> Magnetic core saturation is also a problem with large signals or where a large-amplitude normal operating signal occurs.

Although many problems exist in using filters for EMP protection, if the EMP-induced signal level is known or can be estimated, and if the load and source impedances of the equipment to be protected are known, a filter can be designed to safely attenuate the signal. However, if a large amount of attenuation is required in the same frequency bandwidth as the normal operating signal, normal operation of the circuit can be affected. If a simple filter is adequate, the cost for protection will be small. Where a small amount of attenuation is adequate, ferrite beads provide an inexpensive and simple means of providing protection. Since large signal attenuation

<sup>2</sup>IIT Research Institute, DNA EMP Awareness Course Notes, DASA 01-69-C-0095, Defense Nuclear Agency, DNA 2772T (August 1971), 41.

<sup>16</sup>W. Ricketts et al., *EMP Radiation and Protective Techniques*, John Wiley and Sons (1976), 174-179, 171.

or not knowing the amount of signal attenuation required creates problems of cost, size, and filter design, the use of a nonlinear TPD by limiting and fixing the disturbance signal level reduces the required amount of attenuation and establishes the amount of attenuation required. Since active TPDs have finite initiation times, high-amplitude residual pulses of very short duration can bypass the TPD. A filter can provide a simple and inexpensive means of attenuating these high-frequency pulses. Table 6 provides a summary of the frequency range of various filters and their advantages and disadvantages.<sup>3</sup>

TABLE 6 SUMMARY OF FILTER FREQUENCY RANGE, ADVANTAGES, AND DISADVANTAGES

Filter class	Filter type*	Useful frequency range (Hz)	Significant advantages	Significant disadvantages
Discrete R, L, C	1, 2, 3, 4	to $10^6$	Versatile, low cost	Large for low frequency Low Q
Ferrite beads	1	$10^6$ to $10^8$	Versatile, dissipative with low passband loss	Spurious resonances Saturation
Filter connector	1	$10^4$ to $10^6$	Design integration simplicity Dissipative	Spurious resonances Saturation
Coaxial	1, 2, 3, 4	$10^7$ to $10^9$	High-frequency use, low parasitics	Large size
Crystal	3, 4	$5 \times 10^6$ to $1.5 \times 10^8$	High Q, small size	Spurious resonances, high cost
Ceramic	3	$10^6$ to $10^7$	High Q, small size	Spurious resonances, not IC compatible
Mechanical	3, 4	$0.1$ to $2 \times 10^4$	High Q	Limited range, not IC compatible, high insertion loss

<sup>3</sup>R. A. Perala and T. F. Ezell, *Engineering Design Guidelines for EMP Hardening of Naval Missiles and Airplanes*, Mission Research Corp., prepared for the Naval Ordnance Laboratories under contract N60921-73-C 0033 (December 1973), Ch 7, p16

### 3.4 Digital Circuit Filtering

The disadvantage of using filters to protect digital circuits is that the signal pulse frequency spectrum and the EMP disturbance signal frequency spectrum overlap. Therefore, designing a filter to attenuate the disturbance input will also attenuate the normal operating pulse signals. The Fourier integral representation of a nonperiodic pulse can be shown to be

$$f(t) = \frac{2}{\pi} \int_0^{\omega_0} \frac{\cos \omega t \sin \omega}{\omega} d\omega \quad (21)$$

This expression is integrated to a finite frequency limit  $\omega_0$  instead of infinity. The effect of this is the same as having the pulse pass through an ideal filter with a nondimensional cutoff frequency of  $\omega_0$  rad/unit time. Figure 38 shows the time waveform of a pulse effectively passing through an ideal filter with successively higher nondimensional cutoff frequencies. For the No 1 ESS telephone system, a typical pulse length is 500 ns.<sup>24</sup> Because of the symmetry of the pulse around the time origin, time and frequency are nondimensionalized by 250 ns. Therefore, the cutoff frequencies in figure 38 for a 500-ns pulse are

$$\omega_0 = 8 \frac{\text{rad}}{\text{unit time}} = \frac{8 \text{ rad}}{2.5 \times 10^{-9} \text{ s}} = 3.2 \times 10^7 \text{ rad/s} = 5.1 \text{ MHz} \quad (22)$$

$$\omega_0 = 16 \frac{\text{rad}}{\text{unit time}} = 10.2 \text{ MHz} \quad (23)$$

$$\omega_0 = 32 \frac{\text{rad}}{\text{unit time}} = 20.4 \text{ MHz} \quad (24)$$

Thus, a filter cutting off frequencies above 5.1 MHz severely reduces the pulse risetime (fig. 38). Increasing the cutoff frequency to 10.2 MHz improves the risetime, and increasing the cutoff frequency to 20.4 MHz provides good risetimes. Even if most of the EMP energy coupled to a digital circuit were above 10 MHz, practical filters are not ideal, and producing a large attenuation at 10 MHz requires a cutoff frequency well below 10 MHz. Therefore, with a realistic filter the pulse risetime will be decreased. The effect of placing a filter in a digital circuit has to be determined by experimentation or computer simulation,<sup>7-10</sup> in order to determine whether the circuit will still function properly. Experimentation or computer simulation may also be necessary in order to determine whether the filter can protect the circuit from an EMP induced disturbance. It will be necessary to iterate between observing the effect of the filter on normal circuit operation and the damage protection provided by the filter in order to determine whether a filter can be found to protect the circuit which does not affect normal circuit operation. Because of the bandwidth overlap between the pulse signal (fig. 38 and eqs 22-23) and the EMP signal bandwidth (fig. 39), finding such a filter may not be possible.

<sup>7</sup>Allan F. Malmberg, *NET-2 Network Analysis Program Release 9*, BDM Corporation, prepared for HDL under contract DAAG39-70-C-0050, HDL-050-1 (September 1973)

<sup>8</sup>Laurence Nagel, *SPICE-2: A Computer Program to Simulate Semiconductor Circuits*, Electron Research Laboratory College of Engineering, University of California at Berkeley (May 1976)

<sup>9</sup>L. D. Milliman et al., *CIRCUS: A Digital Analysis Program of Electronic Circuits—Program Manual*, The Boeing Corporation, prepared for HDL under contract DA-49-186-AMC 346 (X), Harry Diamond Laboratories, HDL 346-2 (January 1967)

<sup>10</sup>BDM Corporation, *NET-2 Network Program Analysis Addendum to User's Manual for Release 9.1*, prepared for HDL under contract DAAG39-77-C-0150, BDM/W 77-573 TH (November 1977)

<sup>24</sup>David Talley, *Basic Telephone Switching Systems*, Hayden Book Co (1969)

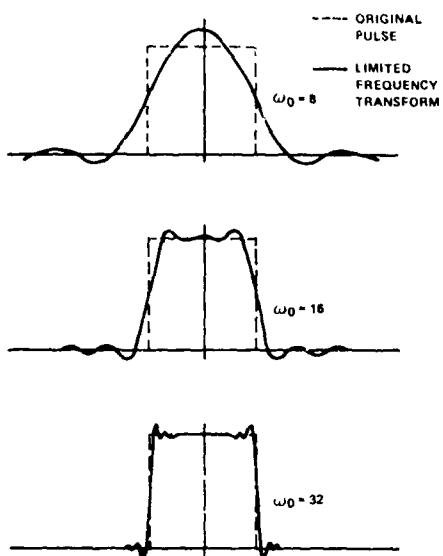


Figure 38 Fourier transform of a pulse showing improved approximation to pulse as frequency is increased

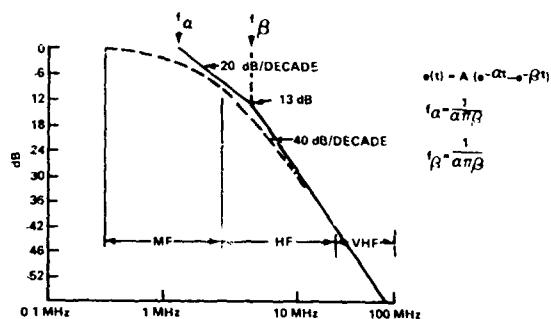


Figure 39 Normalized bandwidth of an EMP signal with an exponential rise and decay

### 3.4.1 Balanced Digital Signal Transmission

As was the case with analog signals (sect. 3.3), a large common-mode signal may be induced on a bundle of wires carrying digital signals. The use of pulse transformers to provide good common-mode rejection has been investigated by Hampel.<sup>25</sup> Two types of transformer circuit arrangements and two types of transformer wiring schemes were studied by Hampel. Figure 40 shows the shunt and series transformer circuit arrangement and figure 20(a) shows the bifilar and

<sup>25</sup>D. Hampel et al, *EMP Hardened Circuits*, RCA, prepared for HDL under contract DAAK02-70-C 0415 (June 1973) AD-911-349

standard transformer wiring schemes that were used. Before investigating the common-mode rejection produced by the different transformer wiring arrangements, the differential pulse capability of the transformers must be determined. In general, large transformers are good for long pulses and small transformers are good for short pulses. It was found that a single, large commercial transformer ( $5 \times 5 \times 3.7$  cm) could transmit from 1 kbs to 10 mbs because of the low leakage inductance and low winding capacitance of the transformer. However, this unit costs \$75 compared to the \$2 to \$3 cost of miniature pulse transformers. Further information on pulse transformer design can be found in Ferroxcube Corp. Bulletin 330.<sup>17</sup>

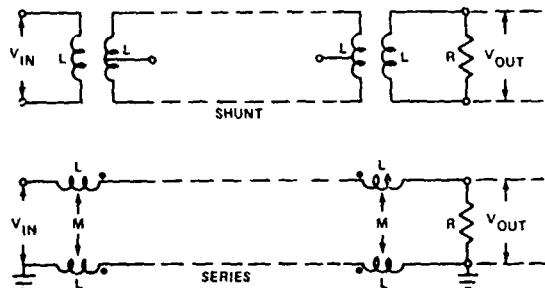


Figure 40 Shunt and series transformer wiring configuration

The series wiring scheme of a transformer (fig. 40) has a differential output voltage given by

$$V_{out} = \frac{RV_{in}}{s(L - \frac{M^2}{L}) + R} \quad (25)$$

where  $L$  = primary inductance, H  
 $M$  = mutual inductance, H  
 $R$  = load resistance, ohms  
 $s$  = Laplace operator  $s^{-1}$

If the transformer coupling is unity and the primary and secondary inductances are equal, then  $L = M$ , and equation 25 becomes

$$V_{out} = V_{in} \quad (26)$$

The common-mode signal produced by an input driving both transmission lines relative to ground (fig. 41) has an output given by

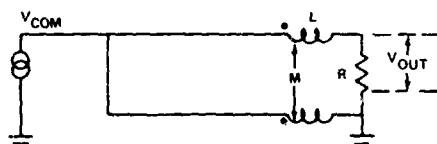


Figure 41 Common-mode driving of series transformer

<sup>17</sup>Ferroxcube Corporation of America, *Applying Ferroxcube Ferrite Cores to the Design of Power Magnetics*, Bulletin 330 (1966)

$$V_{out} = \frac{RV_{com}}{s(L + M) + \frac{RL}{L - M}}, \quad (27)$$

If again  $L = M$ , then equation 27 becomes

$$V_{out} = 0. \quad (28)$$

This is, of course, the ideal condition, which is never met in a real situation. Also, the development of equations 25 and 27 depends on an ideal ground return<sup>25</sup>. If the ground return has either impedance or a signal induced on it, the common-mode rejection will be decreased.

Figure 42 shows the common-mode rejection for transformers with series and shunt circuit wiring and standard and bifilar core winding. Note that the CMR (common-mode rejection) decreases with increasing frequency, except for the bifilar-series winding for which the CMR decreases at both high and low frequency. The bifilar windings provide a better CMR than the standard winding, and the series connected transformer provides a better CMR than the shunt-connected transformer. The series-connected transformer is dependent on a good ground return and may not perform as well in a realistic situation. It does have a response to dc and is therefore better for low-frequency uses. From figure 42 it is seen that either shunt or series wound transformers can provide over 30 dB of CMR for frequencies up to 30 MHz. The energy in an EMP signal is greatly reduced above 30 MHz (fig. 39), so differential signal transmission and isolation transformers can provide adequate EMP protection. Since the transformers must not only reject a common-mode signal but must also pass a differential signal, the differential response of the transformers was also investigated. Figure 43 shows that the bifilar windings provide a wider bandwidth signal transmission than the standard winding for both series- and shunt-connected transformers. The bifilar windings produced a wider bandwidth differential signal and a better CMR than the standard winding.

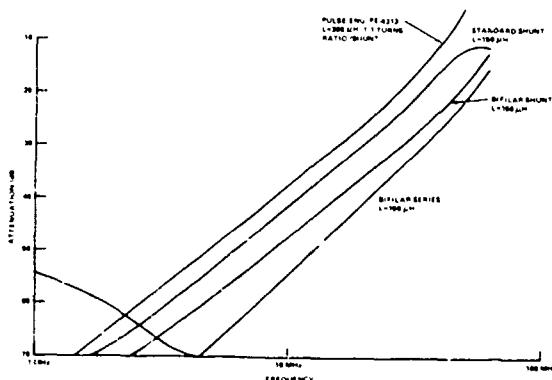


Figure 42 Common-mode rejection test of shunt and series transformer configuration with standard and bifilar windings

<sup>25</sup>D. Hampel et al., *EMP Hardened Circuits*, RCA, prepared for HDL under contract DAAK02-70 C 0415 (June 1973) AD-911 349

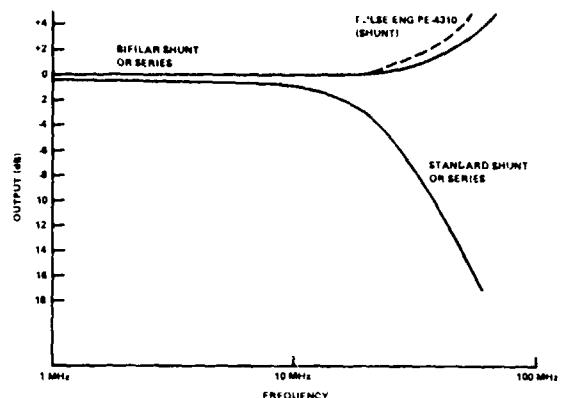


Figure 43 Differential output of transformers with shunt and series configuration and bifilar and standard windings

### 3.4.2 Balanced Circuit Example

Figure 44 shows an isolation transformer used in the No. 1 ESS. A differential signal from the pulse transmitter produces a pulse signal that goes to the receivers. A common-mode signal on the twisted pair produces opposite polarity signals to the pulse receivers. If the transformer primaries have the same inductance and coupling to the secondary, then the common-mode signal is cancelled. Testing the CMR of the 2598 transformer showed a common-mode rejection of 40 dB at 30 MHz.\* A minimum rejection of 20 dB occurs at higher frequencies. EMP pulse simulation testing of the No. 1 ESS at Pickens, MS as part of the PREMPT (Program for EMP Testing) did not show any large signal pickup at the pulse receiver inputs.<sup>26</sup> This demonstrates the effectiveness of isolation transformers in a digital system.

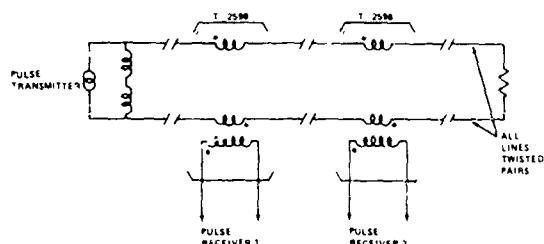


Figure 44 No. 1 ESS pulse transmission system

<sup>26</sup>Joseph R. Miletta et al, Final Report HEMP Assessment of the AUTOVON No. 1 ESS, prepared by Harry Diamond Laboratories for the Defense Communications Engineering Center (September 1971) (SECRET)

\*Informally reported by Christian Fazi and Paul Taltavull

#### LITERATURE CITED

- (1) E. F. Vance et al, Technical Inputs and EMP Design Practices for Intrasite Cabling of Telecommunication Facilities, Stanford Research Institute, prepared for HDL under contract DAAG39-76-C-0021 (June 1977)
- (2) IIT Research Institute, DNA EMP Awareness Course Notes, DASA 01-69-C-0095, Defense Nuclear Agency, DNA 2772T (August 1971)
- (3) R A Perala and T F Ezell, Engineering Design Guidelines for EMP Hardening of Naval Missiles and Airplanes, Mission Research Corp., prepared for the Naval Ordnance Laboratories under contract N60921-73-C-0033 (December 1973)
- (4) M E. Van Valkenburg, Network Analysis, Prentice Hall, Inc (1955).
- (5) Donald G Fink, Standard Handbook for Electrical Engineers, 10th edition, McGraw-Hill (1969)
- (6) Philip R Geffe, Simplified Modern Filter Design, Hayden Book Company, Inc. (1963)
- (7) Allan F Malmberg, NET-2 Network Analysis Program Release 9, BDM Corporation, prepared for Harry Diamond Laboratories under contract DAAG39-70-C-0050, HDL-050-1 (September 1973)
- (8) Laurence Nagel, SPICE-2: A Computer Program to Simulate Semiconductor Circuits, Electron Research Laboratory, College of Engineering University at Berkeley (May 1976)
- (9) L D. Milliman et al, CIRCUS. A Digital Analysis Program of Electronic Circuits—Program Manual, The Boeing Corporation, prepared for HDL under contract DA-49-186-AMC-346(X), Harry Diamond Laboratories, HDL-346-2 (January 1967).
- (10) BDM Corporation, NET-2 Network Program Analysis Addendum to User's Manual for Release 9.1, prepared for Harry Diamond Laboratories under contract DAAG39-77-C-0150, BDM/W-77-573-TR (November 1977)
- (11) H A Lasitter et al, Nuclear Electromagnetic Pulse Protective Measures Applied to a Typical Communication Shelter, Naval Civil Engineering Laboratory (April 1970) AD-707-696.
- (12) Donald L. Chaffee and D. B Clark, Insertion Loss and Pulse Response of Power Line Filters, Naval Civil Engineering Laboratory, Technical Note N-1111 (June 1970).
- (13) R Sherman et al, EMP Engineering and Design Principles, Bell Laboratories, Electrical Protection Department (1975)
- (14) Capcon, Inc , Lossyline EMI Absorptive Filters, Cat. No LSF-6.
- (15) Corcom, Inc., RFI Power Line Filters, Cat No 763A
- (16) L W. Ricketts et al, EMP Radiation and Protective Techniques, John Wiley and Sons (1976)

LITERATURE CITED (Cont'd)

- (17) Ferroxcube Corporation of American, Applying Ferroxcube Ferrite Cores to the Design of Power Magnetics, Bulletin 330 (1966)
- (18) Ferronics, Inc., Wide Band Ferrite Cores, Bulletin 401, Fairport New York (no date)
- (19) D. M. Steward Manufacturing Company, Technical Bulletin Ferrite Shielding Beads, Chattanooga, TN (no date).
- (20) Indiana General, Ferrites Catalog 209, Keasbey New Jersey (no date)
- (21) Cannot ITT, Miniature Rectangular Filter Contact Connectors, DPXJ/DPX\*J (January 1973)
- (22) R. C. Keyser, Satellite Hardening Techniques: Evaluation and Test Definition, IRT Corp., Air Force Weapons Laboratories, AFWL-TR-74-328 (September 1975). AD-B007040L
- (23) Ferroxcube Corporation of America, How to Design Optimum Inductors, Bulletin 220-B (no date)
- (24) David Talley, Basic Telephone Switching Systems, Hayden Book Co (1969)
- (25) D. Hampel et al, EMP Hardened Circuits, RCA, prepared for HDL under contract DAAK02-70-C-0415 (June 1973) AD-911-349
- (26) Joseph R. Miletta et al, Final Report: HFMP Assessment of the AUTOVON No 1 ESS, prepared by Harry Diamond Laboratories for the Defense Communications Engineering Center (September 1971) (SECRET)

DISTRIBUTION

ADMINISTRATOR  
DEFENSE TECHNICAL INFORMATION CENTER  
ATTN DTIC-DDA (12 COPIES)  
CAMERON STATION, BUILDING 5  
ALEXANDRIA, VA 22314

DIRECTOR  
DEFENSE COMMUNICATIONS ENGINEERING CENTER  
ATTN R70/C. STANSBERRY (25 COPIES)  
1860 WIEHLE AVENUE  
RESTON, VA 22090

EMA  
1990 SOUTH GARRISON STREET  
P.O. BOX 26263  
ATTN DR. R. PERALA  
DENVER, CO 80226

TRW  
DEFENSE SPACE SYSTEMS GROUP  
ONE SPACE PARK  
ATTN E. F. CHIVINGTON (2 COPIES)  
REDONDO BEACH, CA 90278

XRI  
1404 EARNSHAW COURT  
ATTN V. MARTINS (2 COPIES)  
RESTON, VA 22090

MR. EDWARD J. VANCE  
ROUTE 3, BOX 268A  
FORT WORTH, TX 76140

DIRECTOR  
DEFENSE NUCLEAR AGENCY  
ATTN: RAEV (2 COPIES)  
WASHINGTON, DC 20305

LAWRENCE LIVERMORE NATIONAL LABORATORIES  
UNIVERSITY OF CALIFORNIA  
P.O. BOX 5505  
ATTN L. MARTIN  
LIVERMORE, CA 94550

US ARMY ELECTRONICS RESEARCH  
& DEVELOPMENT COMMAND  
ATTN TECHNICAL DIRECTOR, DRDEL-CT  
ATTN LEGAL OFFICE

HARRY DIAMOND LABORATORIES  
ATTN CO/TD/TSO/DIVISION DIRECTORS  
ATTN RECORD COPY, 81200  
ATTN HDL LIBRARY, 81100 (3 COPIES)  
ATTN HDL LIBRARY, 81100 (WOODBRIDGE)  
ATTN TECHNICAL REPORTS BRANCH, 81300  
ATTN CHAIRMAN, EDITORIAL COMMITTEE  
ATTN CHIEF, 21000  
ATTN CHIEF, 21100  
ATTN CHIEF, 21200  
ATTN CHIEF, 21300  
ATTN CHIEF, 21400 (4 COPIES)  
ATTN CHIEF, 21500  
ATTN CHIEF, 22000  
ATTN CHIEF, 22100  
ATTN CHIEF, 22300  
ATTN CHIEF, 22800  
ATTN CHIEF, 22900  
ATTN CHIEF, 20240 (3 COPIES)  
ATTN BEILFUSS, J., 21400  
ATTN MILETTA, J., 21100 (10 COPIES)  
ATTN CAMPI, M., 11200  
ATTN LANHAM, C., 00210  
ATTN GOODMAN, R., 34400  
ATTN VALLIN, J., 22100  
ATTN GEIPE, T. M., 22100  
ATTN MEYER, O. L., 22300  
ATTN TUTTLE, J., 21400  
ATTN REYZER, R., 21300  
ATTN FEMANIAS, R., 21100  
ATTN ROFFMAN, G. L., 11400 (8 COPIES)